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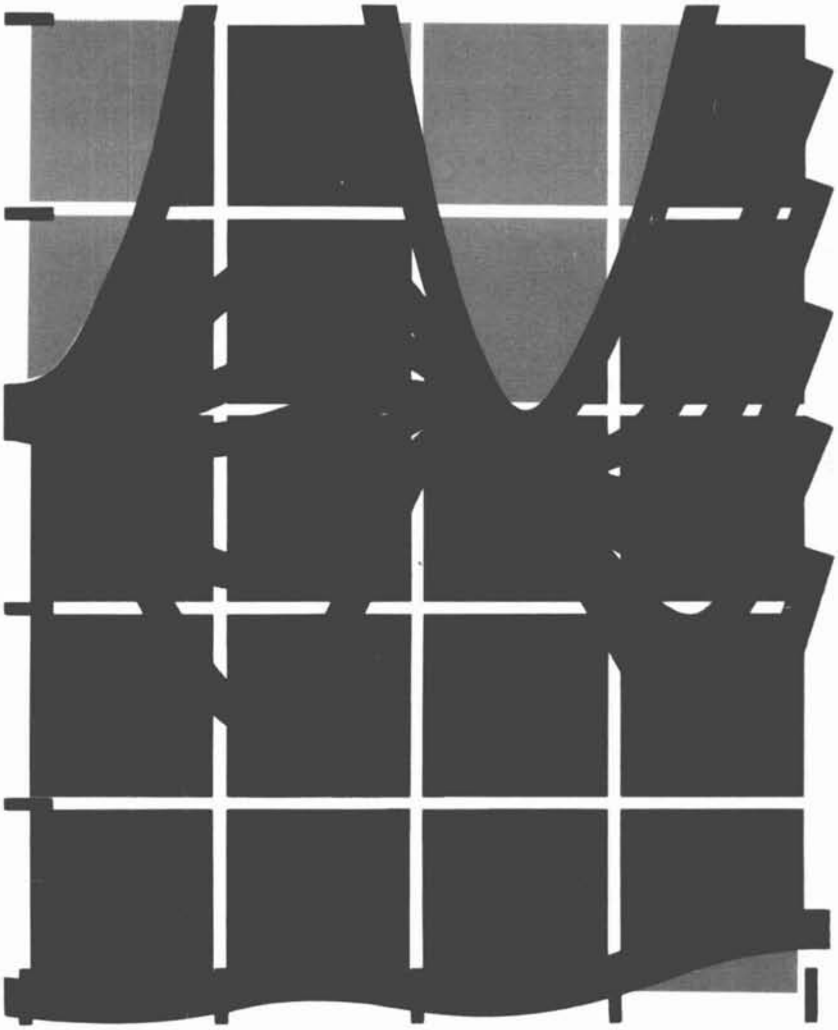
ham radio magazine



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		FREQ	STABILITY-DESIGN	BELOW 500 MHz	ABOVE 500 MHz		12 MHz	60 MHz	MAX FREQ			
K-7000-AC	550 MHz	5.24288 MHz	± 1 PPM - RTXO	15 mV -24 DBM	N/A	(2) .1, 1 SEC	10 Hz	10 Hz	100 Hz	No	No	Yes
7010-S	600 MHz *1 GHz	10 MHz	± 1 PPM - TCXO *±0.1 PPM-OCXO	10 mV -27 DBM	20 mV -21 DBM	(3) .1, 1, 10 SEC	.1 Hz	1 Hz	10 Hz	Yes	No	Yes
8010-S	600 MHz *1 GHz	10 MHz	± 1 PPM - TCXO *±0.05 PPM-OCXO	10 mV -27 DBM	20 mV -21 DBM	(4) .01, .1, 1, 10 SEC	.1 Hz	1 Hz	10 Hz	Yes	Yes	Yes
8013-S	1.3 GHz	10 MHz	± 1 PPM - TCXO *±0.05 PPM-OCXO	10 mV -27 DBM	20 mV -21 DBM	(4) .01, .1, 1, 10 SEC	.1 Hz	1 Hz	10 Hz	Yes	Yes	Yes

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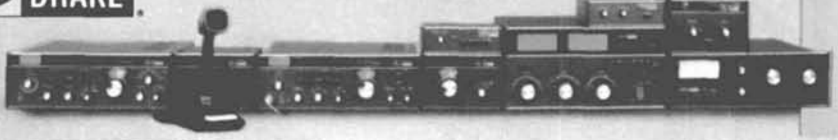

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* Patent pending

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magazine

FEBRUARY 1982

volume 15, number 2

ham radio magazine is published monthly by Communications Technology, Inc
Greenville, New Hampshire 03048-0498
Telephone: 603-878-1441

subscription rates

United States: one year, $16.50
two years, $28.50; three years, $38.50

Canada and other countries (via Surface Mail)
one year, $21.50; two years, $40.00
three years, $57.00

Europe, Japan, Africa (via Air *Forwarding Service*) *one year*, $28.00

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foreign subscription agents

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Cassette tapes of selected articles from *ham radio* are available to the blind and physically handicapped from Recorded Periodicals
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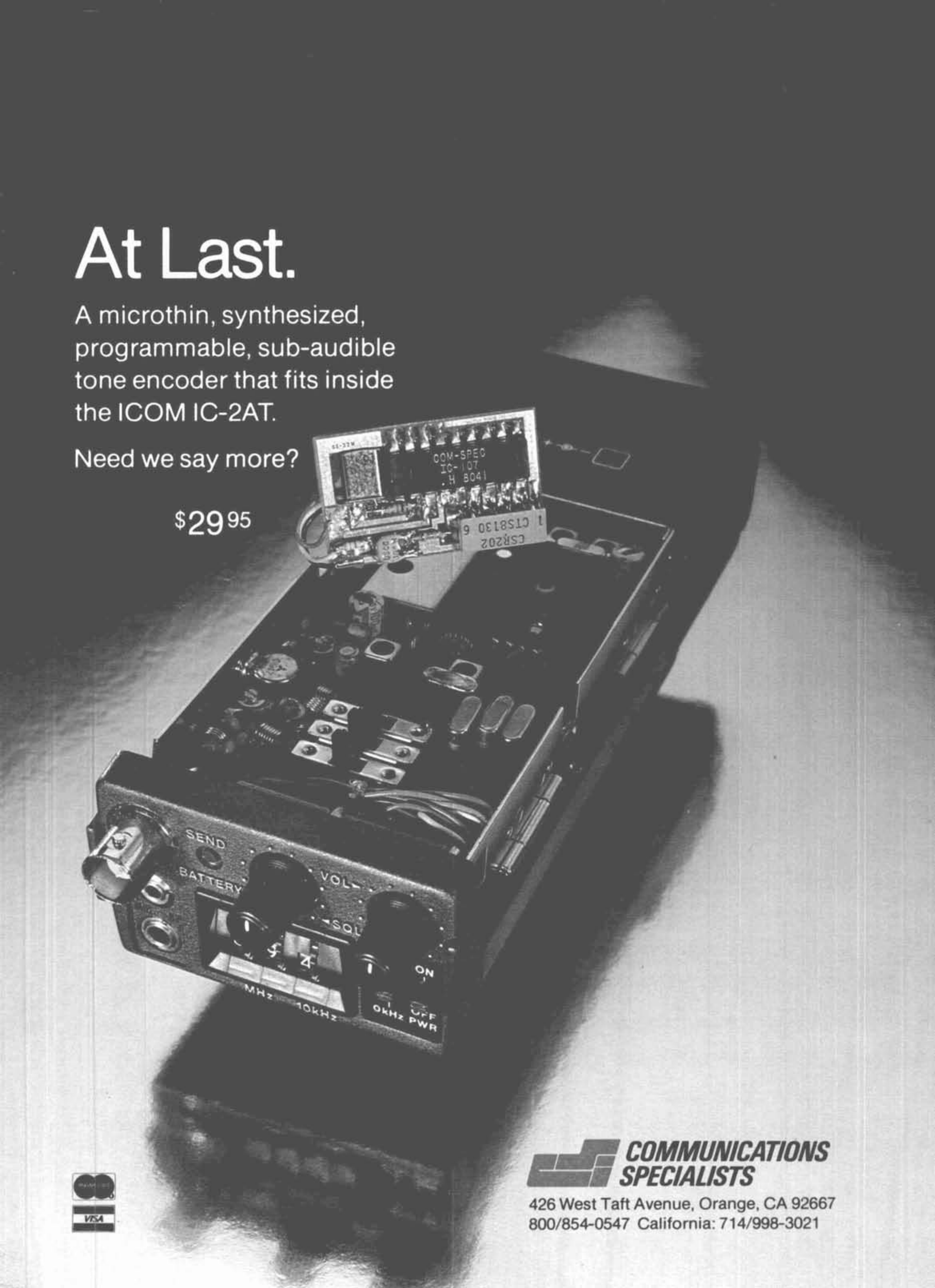
Second-class postage paid at Greenville, N.H. 03048-0498 and at additional mailing offices
ISSN 0148-5989

Postmaster send Form 3579 to *ham radio* Greenville, New Hampshire 03048-0498

contents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Observation & Opinion

This month we present a guest editorial by *ham radio* staff member and assistant publisher J. Craig Clark, N1ACH, who has some comments regarding the condition of the 160-meter band. Alf Wilson, W6NIF, **editor**.

I am more than a little concerned about something that happened recently on 160 meters. It was unnecessary and violated my idea of how one should solve a problem. It also was a display of the sorry state of sportsmanship now more than ever commonplace in Amateur Radio.

What am I talking about? On the weekend of December 4-6, 1981, the annual ARRL 160-meter DX contest was held. This was the first contest run under the ARRL's new voluntary 160-meter band plan, which partitions the band into distinct CW and SSB segments. An East Coast Amateur who was upset with the band plan had set up shop right in the middle of the DX window and was soliciting names for a petition to protest the new plan. He was on frequency for quite a while and kept many hams from working any DX until he went away.

Now let me state that I, too, dislike the band plan. Band plans are great where they are needed. I think that band plans can be of tremendous help on the VHF/UHF bands for accommodating repeaters, weak signal, EME, and other special forms of communications.

On the other hand, the 160-meter band has been called the "gentlemen's band." For more years than I have been a ham, 160 has survived partitioning for LORAN A, low power, poor propagation, and other maladies besides. Both modes of transmission, CW and Phone are used, and only occasionally has an argument erupted. With only 50 kHz of space the band was sometimes very cramped, but always cooperation and peaceful coexistence prevailed. By convention, CW stations congregate in the bottom 10-15 kHz of the band and move higher only when things get crowded. The sidebanders (and a-m'ers, who also, by the way, have every right to be on the air) set up shop above 1.815 MHz. Peaceful and cooperative.

Now, with the elimination of all restrictions below 1.9 MHz, the problems of limited space should be behind us. And until this weekend, things seemed to be OK. But they are not.

Now this East Coast ham who was getting callsigns for his petition went about his project in the wrong way. Instead of interfering during a contest (remember, I don't like the plan either), why not get names over a period of several nights or weeks during casual operating? That, to me, is a far less obnoxious way to go about putting together a protest petition, and one that might get more supporters. By interfering with the contest he may have made more enemies than friends.

If you don't like something, write the ARRL President, W2HD, local ARRL officials, directors, or the author of the band plan, Dave Sumner, K1ZZ. Make them realize that the whole world does not endorse the ARRL plan. Maybe then we can get them to revoke or rethink their plan. Is it really necessary to have more regulations, or could we make do without the plan?

I just wonder — did the directors ever see the comments about the band plan that were sent in? If one looks at the *QST* correspondence column, one would think the whole world went along with the plan. I would be surprised to learn that I'm the only Amateur who was unhappy about the band plan. Last night there were at least five SSB QSOs in progress below 1.825 MHz. Either they never heard of the plan or they also don't agree with what has been adopted.

In this day and age of deregulation, why not let the 160-meter band be an example of how peaceful coexistence can work? It already has worked for many years. Is there reason to expect that it can't now? I don't think so. I don't buy the arguments about technical limitations or whatever as being sound reasons for a band plan. I am afraid it is needless regulation and I would rather see no plan than this one.

I support action to revoke the ARRL band plan. But I will not ruin the pleasure and operating enjoyment of others to show my sentiments. I am afraid that this controversy may turn 160 meters from a "gentlemen's band" into a mess the likes of 20 meters. Is that what we want? I don't think so. For those who don't like the band plan, let's not ruin one band where we all can have fun. Let's all observe the DX window — it was there long before the band plan. And let's make sure ARRL officials know of our feelings. Begrudgingly, I will abide by the band plan. I will, however, continue to urge the ARRL President and directors to reconsider it.

J. Craig Clark, N1ACH, assistant publisher

comments

9.3 MHz anyone?

Dear HR:

I have several comments pertinent to Gary O'Neil's fine article on antenna traps.[1] Although I appreciate his referencing an article of mine,[2] I must point out that "my" method of trap construction is not original with me. I can recall seeing more than one mention of the use of coax as a capacitor prior to my application, which was stimulated by a clever technique developed by Mathison.[3]

It is very important to realize that there is no such thing as "1.25-inch pipe"! There is 1¼-inch pipe, which actually is 1.66 inches in diameter, and there is 1.25-inch tubing, which actually is 1.25 inches in diameter. *ham radio*'s policy of printing all non-integral quantities as decimals is very confusing in this case: author O'Neil means *1¼-inch pipe* (1.66-inch diameter), but *ham radio* printed "1.25-inch PVC stock" on page 14. There's a big difference, as I found out when I constructed 40-meter O'Neil traps on 1.25-inch-diameter forms: Anybody need a set of 9.3 MHz traps?

In a similar vein, it also is important to note that *any* length of wire connected to a tuned circuit such as an antenna trap will cause its resonant frequency, when measured on the bench with a dip meter, to appear to be lower than it actually is. The 3-inch connecting leads specified on page 14 will cause the resonant frequency of a 7-MHz trap, as measured with a dip meter, to appear to be 30 to 45 kHz lower than the true value. This may be verified by connecting leads of differing lengths, measuring the apparent resonant frequency each time, plotting these frequencies as a function of lead length, and finally extrapolating backwards to zero lead length. This characteristic in no way affects the value or utility of O'Neil's method, but it is a potential source of trouble for those attempting to build and tune traps.

I have always understood that PVC is a poor choice as a coil form because of its high rf loss.[4] I have no feel for how serious this might be in the present application, but comments from those who are knowledgeable would be useful. If PVC truly is a very lossy material in the high-frequency region, it probably should not be used in antenna traps.

Finally, it is interesting to compare O'Neil's coax-cable trap design to that of Johns.[5] It might be useful to perform a quantitative comparative study of the two designs.

Gary O'Neil is to be complimented on his thoroughness and ingenuity in this project. The resulting article is interesting, useful, and well written.

references

1. Gary E. O'Neil, "Trapping the Mysteries of Trapped Antennas," *ham radio*, October, 1981, page 10.
2. Gary E. Myers, "A Two-Band Half Sloper Antenna," *QST*, June, 1980, page 32.
3. J.R. Mathison, "Inexpensive Traps for Wire Antennas," *QST*, February, 1977, page 18.
4. W1CKK, "A Dielectric No-No," Hints & Kinks for the Radio Amateur, 1978 Edition, ARRL, page 68.
5. Robert H. Johns, "Coaxial Cable Antenna Traps," *QST*, May, 1981, page 15.

Gary E. Myers, K9CZB
Naperville, Illinois

satellite locator

Dear HR:

The satellite locator program for the TI59 by Mr. Walter Pfiester, W2TQK, (October, *ham radio*) works OK for azimuth but needs the following change or addition to run for elevation.

In program step 25 (STO, 03, R/S), insert (between R/S and RCL, 05. It then looks like this: STO, 03, R/S, (, RCL, 05.

I made one other change at program steps 21, 22, and 23. Change them to INV, TAN, +, *RCL, 10,* =, STO, 03. This way I can enter 180 or 360 as required at register 10 when I key in the other data. A *very good job*, Walt.

Frank Sheehan, W1DHX
Port Isabel, Texas

line losses

Dear HR:

With respect to the recent letters by WD4KMP and WB9TQG in the September, 1981, issue, I don't think it is so difficult to demonstrate that line losses are higher when the line has a high SWR. This is not a mathematical proof, but rather a heuristic, common-sense type.

If a line has a high SWR, it has, by definition, peaks and valleys of voltage and current on it. Since the power lost in the line is divided into two major parts, the loss in the dielectric and in the copper, and since one varies with the square of the voltage while the other varies with the square of the current, it follows naturally that the peaks of voltage and current increase the line losses. It is true that the valleys decrease the losses, but since the power losses vary as the square of the voltage and current, the added losses more than make up for the reduced losses at the valleys. In other words, the least possible loss is in a line that is "flat," where there are no high peaks of voltage or current.

The net result of this is that I agree with WB9TQG and the various references he quotes on line losses.

James N. Thurston, W4PPB
Clemson, South Carolina

upgrade

Dear HR:

I do appreciate Robert Shrader's articles on upgrading. Please keep them coming. Many of us need this approach to make us more proficient.

Lloyd Morse, WA7YJF
Allyn, Washington

More Details? CHECK—OFF Page 100

presstop

SIX RUSSIAN SATELLITES, RS3 through RS8, were launched on December 17 and are now in a nearly circular orbit around the earth at an average altitude of nearly 1700 km. The six are steadily moving away from each other with slightly differing orbits, and by December 28 their equatorial crossing times were spread over more than an hour and crossing points nearly 20 degrees.

All Six Have Been Transmitting telemetry data, with each series preceded by the spacecraft's call (e.g., "RS3"). RS3, 5, and 7 all have "robot transponders," and at least one has been worked by a number of stations around the world. Robot availability is indicated by a "CQ," stopping when a signal appears in its input passband. Sending (for example) "RS5 de W9XXX AR" should bring the response "W9XXX de RS5 QSO NR xxx". It may also respond "QRZ," "QRM," or "RPT" if it misses a call, or "QRQ" or "QRS" to calls made below or above its 10-25 WPM acceptance range.

Beacon Frequencies for the even numbered birds are: RS4, 29360/29403; RS6, 29411/29453; and RS8, 29461/29502. Their 40-kHz-wide OSCAR-style transponders have apparently not been activated as of this writing. One indication of transponder status in any of the six is the first, or "K," group telemetry number, which indicates power output. A reading of anything other than "KØØ" should mean the transponder is on.

Interference To The RS Satellites from terrestrial stations is becoming a real problem, with their covering so much of the 29.3-29.5 MHz spectrum. SSB, AM, and FM signals have all been heard in recent weeks on top of or breaking over onto the new satellites. Non-satellite users should try to stay below 29.3 or above 29.5 to avoid the problem.

10 MHZ WAS STILL WELL OFF in the future, so far as U.S. Amateurs are concerned, despite some last-minute efforts by the League and others to push some kind of temporary authorization through by January 1. Though the idea of making the band available for U.S. Amateurs by the January 1st date that was made possible by WARC '79 did have support from several key parts of the Commission, some questions about details prevented complete acceptance. Another difficulty was opposition among current U.S. users of the 10.10-10.150 MHz segment, both common carrier and government. As a result, barring a drastic change in the situation as it exists at press time, it's unlikely that the band will be opened to U.S. Amateurs until all treaty and rule making procedures are accomplished.

Considerable 10-MHz Activity was expected from other parts of the world on January 1, however. West Germany, Switzerland, and Australia are among the major nations joining the British Isles on the band, and at least one Scandinavian country and some smaller nations were expected on as well.

The "CW Only" Concept for the new band seems to be meeting some resistance in IARU Region 3. With the exception of Japan, the Amateur population of the Far East is quite sparse. As a result, Amateurs there (and possibly some government officials) are pushing for some sort of phone sub-band. How that could be accomplished in a 50-kHz wide band is another question.

JAPANESE REPEATERS ARE EXPECTED to be authorized in the near future, and the JARL has announced new VHF/UHF bandplans set up to accommodate them. Key points of the new bandplans are enlarging of the FM segments of the 144 and 440 MHz bands to provide room for repeaters, and moving the present 52.5-MHz beacon frequency to 50.010 MHz. The latter move puts Japan into agreement with the rest of the world, with beacons at the lower band edge to better signal band openings.

RFI CONTROL WON'T BE A PART of K7UGA's pro-Amateur Radio bill as it's shaping up in the House of Representatives. The House version of S.929, termed "Track 1" by subcommittee chairman Timothy Wirth, has deleted both the controversial anti-RFI and CB delicensing provisions. The RFI aspects were strongly opposed by the Electronic Industries Association, while many CB user groups are fighting the delicensing idea. Both items are to be considered later, however, as "Tracks 2 and 3."

FCC Participation In Conventions and similar activities is encouraged by a new provision to the bill added by the House. It would permit the FCC to accept compensation for its attendance at such events, so FCC-administered exams and seminars at hamfests would no longer have to be financed from the commission's budget.

Passage Of This Far-Reaching legislation could take place fairly soon if the Senate concurs with the changes that were incorporated by the House.

EXPANSION OF U.S. PHONE BANDS, 220-MHz phone privileges for Novices, and proposals to change some power limits are all likely to be seeing FCC action in the near future. The phone band expansion, despite opposition from outside the country, will probably come up first. The power limit proposals, one to reduce CW power and the other to permit higher power levels for moonbounce work, are likely to be tied in with long-standing FCC attempts to develop new power-specifying techniques.

BILL SCARBOROUGH, N1BXG, HAM RADIO graphic production manager, has been promoted to assistant advertising manager in addition to his previous responsibilities.

response of pi, pi-L and tandem quarter-wave-line matching networks

Bandwidth considerations using the HP-34C to solve network equations

It is usually assumed — but not shown — that the response of pi and pi-L impedance-matching networks used in exciters and linear amplifiers resembles that of ordinary parallel-resonant LC circuits. The reason is that considerable computation is involved to generate the response curves. With the availability of relatively low-cost programmable calculators such as the HP-34C the problem becomes much simpler. This article presents the response characteristics of several commonly used pi and pi-L networks, developed with the aid of an HP-34C program and the SOLVE capability of this machine for finding specific maxima, minima, and points on the curves. The last part shows the response for some broadband matching systems using tandem transmission lines, each a quarter-wavelength long at the design (center) frequency.

The networks selected are tabulated in **table 1**, which gives the reactance values in ohms for five different input resistances at resonance (matched conditions) and for two different Q values. The relationships used were programmed into the HP-34C; the programs will be found in *The Match Book*, a manuscript now awaiting publication.* The design equations were given in an article by the author some years ago.[1]

The pi-L network assumes the same transformation ratio for the pi and for the output L, as is customary; that is, the input pi matches to the geometric mean ($\sqrt{R_{in} \times 50}$ ohms in this case) between the end resistances, and the L takes it the rest of the way between this intermediate resistance and 50 ohms. Only networks with shunt capacitance and series inductance are considered here, because they have the best harmonic attenuation and are the most com-

*Programs for pi and pi-L networks are available from *ham radio* upon receipt of a self-addressed stamped envelope.

By **R.W. Johnson, W6MUR**, 2820 Grant Street, Concord, California 94520

monly used. The output capacitance of the pi is then combined with the input capacitance of the L to result in a four-reactance network as shown in **table 1**.

network parameters

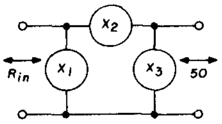
Use of the polar-to-rectangular (and vice versa) *capabilities* of the HP-34C, and its ability to handle subroutines makes it feasible to construct a fairly simple program for finding the input impedance and all of its components: resistance, reactance, impedance and phase angle. **Fig. 1** shows the input impedance of the pi ($Q = 10$) matching 3000 ohms to 50 ohms, and defines the various critical maxima and frequencies:

R_m	maximum resistive component
X_{m1}	maximum positive reactance component
X_{m2}	maximum negative reactance component
$\lvert Z_m \rvert$	maximum magnitude of input impedance
ρ_1	frequency for X_{m1}
ρ_2	frequency for X_{m2}
ρ_3	frequency for R_m
ρ_4, ρ_5	frequencies for $0.707R_m$
ρ_6	frequency for $\lvert Z_m \rvert$
ρ_7, ρ_8	frequencies for $0.707\lvert Z_m \rvert$

table 1. Matching networks studied (all match to 50 ohms).

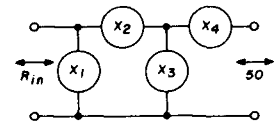
pi network values (ohms)

Q	R_{in}	X_1	X_2	X_3
8	3000	−375.00	382.5542	−173.2051
8	2750	−343.75	356.5016	−117.2604
8	2500	−312.50	328.7586	−91.2871
8	2250	−281.25	300.0000	−75.0000
8	2000	−250.00	270.4791	−63.2456
10	3000	−300.00	321.5834	−60.4858
10	2750	−275.00	297.1778	−54.6729
10	2500	−250.00	272.5235	−49.5074
10	2250	−225.00	247.6236	−44.8211
10	2000	−200.00	222.4734	−40.4888



pi-L network values (ohms)

Q	R_{in}	X_1	X_2	X_3	X_4
8	3000	−375.00	494.7104	−72.8549	129.8650
8	2750	−343.75	456.3520	−69.7073	126.6511
8	2500	−312.50	417.7783	−66.3802	123.1977
8	2250	−281.25	378.9629	−62.8462	119.4592
8	2000	−250.00	339.8733	−59.0706	115.3750
10	3000	−300.00	400.0911	−63.8365	129.8650
10	2750	−275.00	368.9983	−60.9351	126.6511
10	2500	−250.00	337.7391	−57.8778	123.1977
10	2250	−225.00	306.2931	−54.6418	119.4592
10	2000	−200.00	274.6345	−51.1982	115.3750



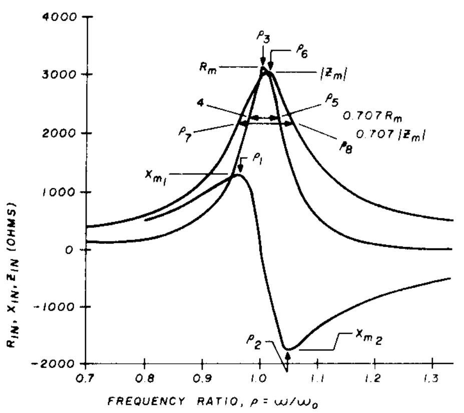


fig. 1. Input impedance of a pi network for matching 3000 to 50 ohms; $Q = 10$.

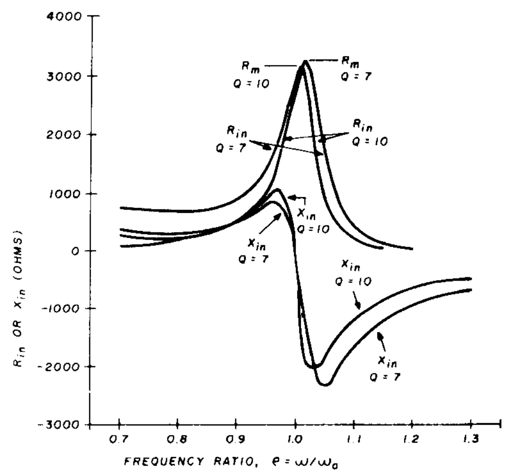


fig. 2. Pi-L network response for matching 3000 to 50 ohms, $Q = 7$ and $Q = 10$.

Table 2 is a tabulation of each of these critical parameters for twenty different networks, covering five different input (design) resistances and two design Q values for each of the pi and pi-L. The maxima were discovered using the SOLVE program in the HP-34C, which requires a progressively refined estimate of what the maximum is so it can be subtracted from the value found by the calculator. This iteration takes several minutes to run for each data point. Since the frequency differences are small, frequency is carried out to six decimal places.

comparison of pi and pi-L response

Fig. 2 is the input impedance response of the pi-L (3000-ohm design) for two different Q values. In this

table 2. Critical frequencies and peak values for typical pi and pi-L matching networks of table 1.

Q	R_{in}	ρ_1	ρ_2	X_{M1}	X_{M2}	ρ_3	R_M	ρ_4	ρ_5	ρ_6	$\|Z_M\|$	ρ_7	ρ_8
						pi network							
8	3000	.93962	1.06433	1299.530	−1702.240	1.004202	3013.515	.963027	1.043497	1.008310	3026.807	.947860	1.072485
8	2750	.93994	1.06399	1177.944	−1575.121	1.004500	2764.344	.963411	1.043451	1.008890	2778.427	.948784	1.072657
8	2500	.94041	1.06352	1058.528	−1445.347	1.004785	2514.969	.963883	1.043306	1.009439	2529.633	.949836	1.072652
8	2250	.94102	1.06292	941.432	−1312.838	1.004500	2265.148	.964434	1.043070	1.009954	2280.321	.951006	1.072473
8	2000	.94184	1.06214	826.759	−1177.570	1.005306	2015.392	.965085	1.042721	1.010426	2030.391	.952296	1.072109
10	3000	.95292	1.04961	1288.648	−1715.496	1.003452	3015.191	.971367	1.033700	1.006824	3030.148	.959790	1.056201
10	2750	.95314	1.04922	1174.164	−1579.818	1.003550	2764.967	.971681	1.033508	1.007011	2779.655	.960394	1.055977
10	2500	.95380	1.04879	1060.995	−1442.766	1.003641	2514.579	.972027	1.033279	1.007185	2528.870	.961041	1.055690
10	2250	.95430	1.04830	949.126	−1304.353	1.003725	2264.024	.972410	1.033007	1.007342	2277.750	.961735	1.055300
10	2000	.95495	1.04774	838.573	−1164.591	1.003800	2013.290	.962836	1.032685	1.007480	2026.777	.962487	1.054884
						pi-L network							
8	3000	.964044	1.043542	885.242	−2153.591	1.008705	3134.395	.980340	1.032437	1.026992	3132.000	.976305	1.060246
8	2750	.964018	1.043432	819.624	−1965.628	1.008562	2869.704	.980291	1.032294	1.016151	2974.000	.978728	1.056761
8	2500	.963500	1.043313	752.888	−1778.753	1.008409	2605.498	.980244	1.032137	1.015433	2698.014	.978530	1.056499
8	2250	.963500	1.043175	685.114	−1593.042	1.008245	2371.818	.980199	1.031963	1.015174	2422.701	.978319	1.056213
8	2000	.963500	1.043016	616.160	−1408.571	1.008072	2078.692	.980158	1.031767	1.014889	2148.336	.978100	1.055887
10	3000	.969708	1.034716	1020.733	−2001.451	1.005421	3080.231	.982849	1.025118	1.010236	3153.783	.979644	1.043867
10	2750	.969733	1.034634	941.466	−1828.769	1.005338	2821.651	.982838	1.025019	1.010930	2887.493	.979544	1.043685
10	2500	.969767	1.034536	861.456	−1656.840	1.005250	2563.337	.982831	1.024909	1.009944	2621.684	.979443	1.043482
10	2250	.969814	1.034500	780.659	−1485.705	1.005155	2305.296	.982829	1.024785	1.009780	2356.375	.979341	1.043259
10	2000	.969800	1.034292	699.026	−1315.429	1.005051	2047.700	.982836	1.024642	1.009602	2091.599	.979240	1.042995

case, the impedance curve, $(|Z|)$, was omitted to avoid cluttering the curves.

Comparison of **fig. 2** with **fig. 1** shows several interesting things. First, although the reactance peaks below and above resonance are not equal in either case, the difference is more marked for the pi-L, for the same design Q. Second, the effective Q of the network, which may be taken as the reciprocal of the difference between ρ_5 and ρ_4 or, if you prefer, between ρ_8 and ρ_7, is considerably higher for the same design Q in the pi-L than for the pi. One would expect this, of course, since there are in effect two additional reactances in the pi-L, and it is well known that the harmonic attenuation is much greater with the pi-L than the pi. However, the same effective Q as for the pi can be obtained at a lower design Q for the pi-L. Instead of using $Q = 10$ as is customary, $Q = 8$ or even $Q = 7$ can be used for the pi-L, and thus save something on the size of the high-voltage input capacitor at the lower frequencies. This is, however, at the expense of more inductance and hence larger coils.

pi-L input impedance near resonance

It is also interesting to examine the input impedance over a very narrow band in the vicinity of resonance. **Fig. 3** is a plot of the *change* in input resistance versus 3000 ohms, and of the total reactance, for the $Q = 10$, 3000/50 ohm pi-L. Shown on the figure is the voice bandwidth for 3.8 kHz at 3.8 MHz, on the assumption that the band is centered at the design frequency where there is zero reactance. It can be seen that there is a total change of about 56 ohms (out of 3000 ohms) over the band due to the fact that R_{in} peaks (in this case) about 20.6 kHz higher than the design frequency. This approximate 2 percent change in resistance presented to the final amplifier tube over the voice bandwidth no doubt contributes something to distortion, although it is probably negligible. In any event, using a lower design Q will certainly correct whatever problem may exist.

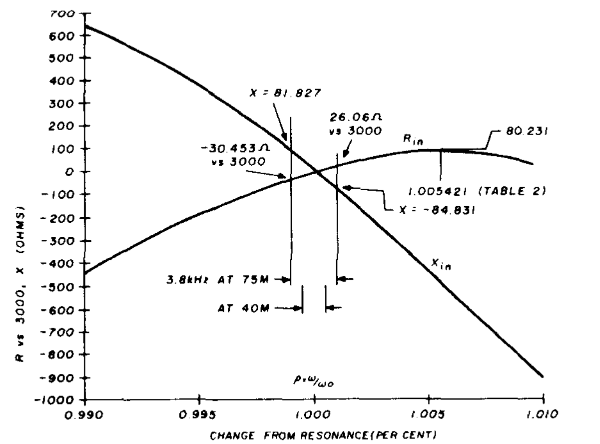


fig. 3. Change in input resistance versus 3000 ohms R_{in} for a Pi-L network over a 3.8-kHz voice bandwidth; $Q = 10$.

Finally, keep in mind that these curves assume a *constant* 50-ohm resistive load on the output. A reactive load with high VSWR will cause the networks to behave quite differently.

While there are no earth-shaking conclusions from all of this, it is nice to know for sure that the input impedance of pi and pi-L networks is indeed very similar to that for parallel-resonant circuits, and the results certainly illustrate very well the powerful capabilities of the HP-34C programmable calculator.

tandem quarter-wave lines

Tandem quarter-wave matching sections, with different Z_0s for each section, tend to exhibit broader bandwidth matching characteristics than most other types. Again, the HP-34C programmable calculator is of enormous value to study some typical cases. The input impedance of a section of transmission line θ degrees long terminated in a complex series impedance $\alpha + j\beta$, with the Z_0 of the line being K times the Z_0 of the previous section is given by

$$Z = K^n Z_0 \left[\frac{\dfrac{Z_L}{K^n Z_0} + jT}{1 + jT \dfrac{Z_L}{K^n Z_0}} \right]$$

$$= K^n Z_0 \left[\frac{\dfrac{\alpha}{K^n Z_0} + j\left(T + \dfrac{\beta}{K^n Z_0}\right)}{\left(1 - \dfrac{\beta T}{K^n Z_0}\right) + j\left(\dfrac{\alpha T}{K^n Z_0}\right)} \right], \quad T = \tan\theta \qquad \textbf{(1)}$$

Introducing the variables $v = \dfrac{K^n Z_0}{\alpha}$ and $w = \beta/\alpha$, **eq. 1** can be rewritten into a form more suitable for calculator use as

$$Z = K^n Z_0 \frac{1 + j(Tv + w)}{(v - wT) + jT} \qquad \textbf{(2)}$$

Expressing this in the series form $\alpha + j\beta$, this is then the load for the next section of line for which the Z_0 is $(K)^{n+1}Z_0$. The HP-34C contains a built-in incrementing program, using the I register, for either increasing or decreasing n. Thus the program is developed starting with $n = 0$ $(K^0 = 1)$, computing the input impedance of this section for which $\beta = 0$ and $\alpha = R$. The resulting values of α and β are stored,

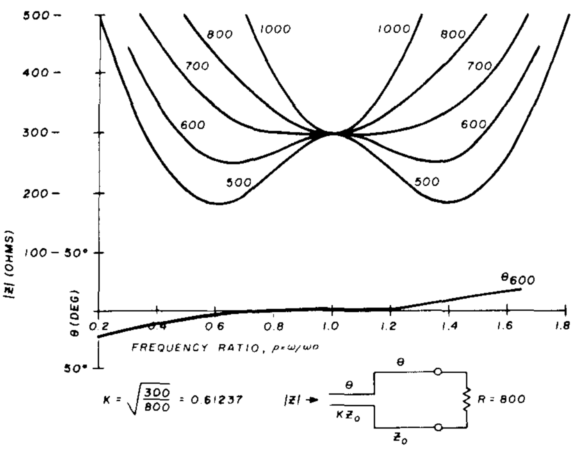


fig. 4. Input impedance of two tandem quarter-wave-length lines for matching 800 to 300 ohms.

then used in the next cycle for which $n = 1$, and so on for however many sections are assumed.

bandwidth considerations

Using this program, several interesting tandem-quarter-wave line cases were examined. **Fig. 4** shows the input impedance of a two-section line for matching 800 to 300 ohms, for five different Z_0 values. There is an "optimum" value of Z_0, in this case around 600 ohms, which will produce the broadest bandwidth. The input phase angle is shown in **fig. 4** for this "optimum" value of Z_0. The practical manifestation of this case would be the first line (connected to the 800-ohm load) to have $Z_0 = 600$ *ohms*, and the second line would have a Z_0 of 367 ohms. The bandwidth would be from 0.38 to 1.62 for VSWR 1.2:1 or less, or a frequency ratio of 4.26:1.

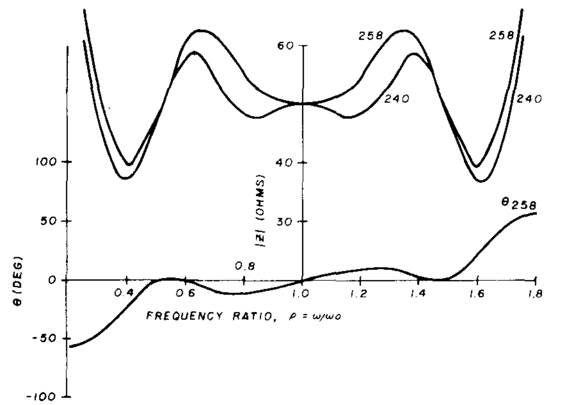


fig. 5. Response of a four-section tandem line, matching 300 to 50 ohms. Each line is nominally one-quarter wavelength long at $\rho = 1$ **for** $Z_0 = 240$ *and 258 ohms.*

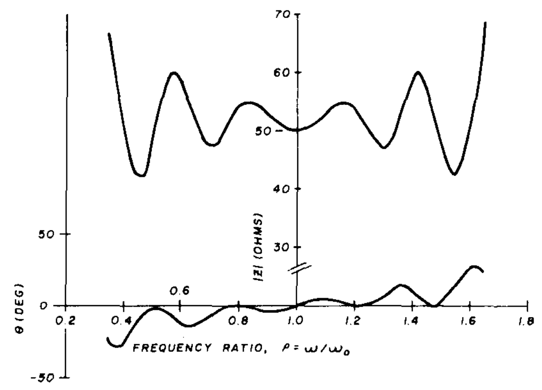


fig. 6. Eight-section quarter-wavelength line showing response for matching 800 to 50 ohms; $Z_0 = 700$ *ohms.*

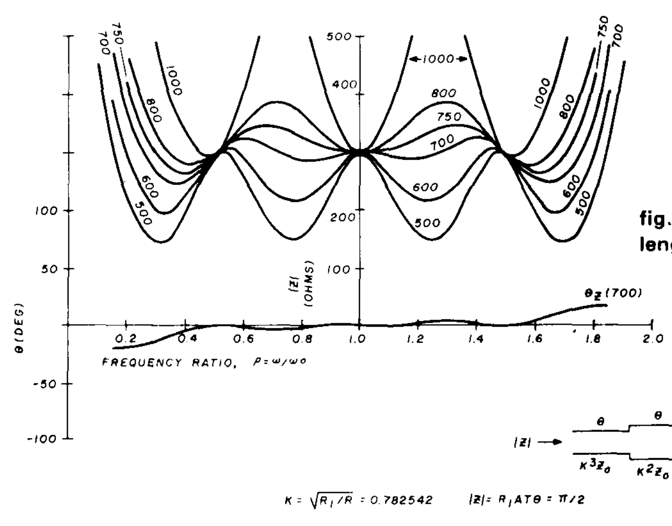


fig. 7. Input impedance of four tandem quarter-wave-length lines for matching 800 to 300 ohms.

As with any matching system, when the transformation ratio increases, the bandwidth decreases. So for a 300:50 ohm case two sections are not enough for a really broad bandwidth. **Fig. 5** shows the results for a four-section line, where the "optimum" Z_0 is now about 258 ohms. Here, the input impedance is within 1.25:1 of 50 ohms from frequency 0.31 to 1.74, a frequency ratio of 5.61:1. The Z_0s would be 258, 165, 105, and 67 ohms for the four sections. The total line length would, of course, be a full wavelength at the center frequency.

Fig. 6 shows the 800:50 ohm case, this time with a still higher transformation ratio. Here, eight sections are necessary, and the optimum Z_0 is about 700 ohms. It can be seen that the bandwidth extends from 0.37 to 1.63, or over a 4.4:1 frequency band within VSWR 1.2:1. Here, the Z_0s would be 700, 495, 350, 247.5, 175, 124, 87.5, and 62 ohms for the eight sections.

Finally, for the 800:300 ohm case, which had a bandwidth of 4.26:1 in a two-section configuration, *ρ can be pushed to 0.2 versus 1.8, or 9:1 in frequency* by going to a four-section line. This is shown in **fig. 7**, in which several Z_0 values have been included so that the reader can see what happens to the various peaks and valleys as Z_0 is changed. The "optimum" case is at $Z_0 = 700$ ohms, the lines being 700, 548, 429, and 335 ohms Z_0 respectively. The phase angle of the input impedance is shown for this optimum Z_0.

Eventually, of course, as one goes to a very large number of sections and a very gradual taper on Z_0, one approaches the tapered transmission line, or exponential line, which has been around for many years as a broadband matching system. Such lines are, however, hard to build and require some large spacings at one end and impracticably small spacings at the other, and may be several wavelengths long. The tandem quarter-wave sections are shorter, more practicable to build, and nearly as good on bandwidth.

summary

This article has attempted to shed some light on bandwidth considerations in pi, pi-L and tandem-line matching systems. Intractable analytically, these various configurations are easily studied with the aid of the enormous computing power of the HP-34C as a relatively inexpensive machine. More information on these and related subjects will be found in *The Match Book*, which some astute publisher will probably produce within a year.

reference

1. R.W. Johnson, "Pi Network Nomograph," *Electronics*, September 12, 1958, page 108.

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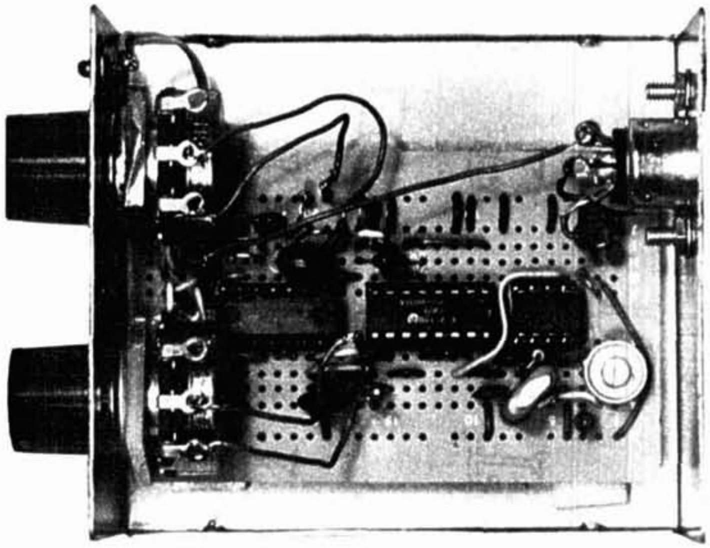

blanking the Woodpecker

part two: a practical circuit

In part one, I discussed a number of aspects of the Woodpecker transmissions which might allow one to go about blanking them. These are, briefly:

1. Repetitive pulses of relatively long duration.
2. Broad bandwidth — typically 50 kHz.
3. Likelihood of occurrence throughout the high-frequency band.
4. Very precise pulse repetition frequency — usually 10 Hz.

Most existing Amateur transceiver noise blankers are designed to deal with intense, short-duration noise spikes of the type generated by car ignitions. These are typically 0.5 millisecond or less in duration. The Woodpecker pulses, on the other hand, are usually 10 to 20 milliseconds long, and consist of a number of spikes of varying amplitude. Because the Woodpecker pulses are quite different from ignition noise, many Amateur transceiver noise blankers do not perform effectively on the pulses, taking only 2 or 3 S-points off. When the Woodpecker is at strength S9 + 20, this is not much help. Some of the newer transceivers (such as the ICOM 720A) are reportedly more effective, but the majority of transceivers in operation today are wiped out by the Woodpecker when it is really strong.

First published in *Amateur Radio Action*, Vol. 4, 1981, Issue 4, P.O. Box 628E, Melbourne 3301 Australia. Reprinted with permission.

As discussed in the first article, conventional noise blankers operate by amplifying noise pulses, and use the result to blank a noise gate early in the i-f chain. This works well if the noise pulse is strong enough to allow the noise amplifier to create a blanking control pulse. However, many of the component spikes in a Woodpecker pulse are not intense enough to trigger the blanking, so a lot of the pulse gets through.

An alternative approach was outlined in my previous article, which has been found to be quite effective against the Woodpecker. This approach makes use of the fact that the Woodpecker pulse repetition frequency (PRF) is precisely defined. About 90 percent of the time it is 10 Hz, to an accuracy of one part in 100,000 or better. Other equally precise PRFs are used, particularly 16 Hz, for the remaining 10 percent or so of the time. This precision in the Woodpecker PRF has led to the idea of a synchronous noise blanker, where a crystal-locked 10-Hz signal is generated locally, synchronized with the Woodpecker, and used to control a blanking gate in the usual way.

In this article, a circuit is described that blanks the Woodpecker pulses early in the receiver i-f stages, before they swamp the AGC and before they get

By David Nicholls, VK1DN, 61 Wybalena Grove, Cook, ACT2614, Australia

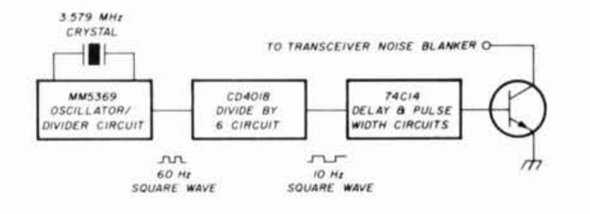

fig. 1. Block diagram of a synchronous noise blanker circuit. (See appendix for notes on devices.)

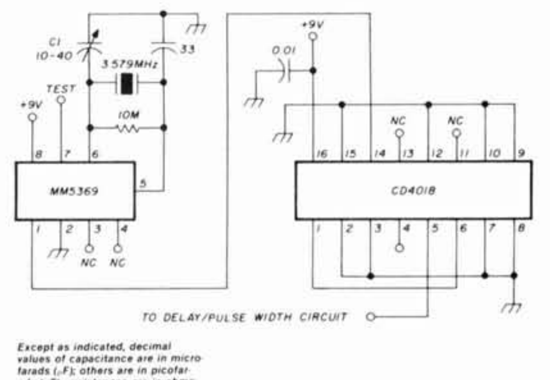

fig. 2. Oscillator/divider circuit.

broadened by the narrow selectivity stages. The circuit is designed to be connected into transceiver noise blankers of the type that operate by reverse biasing diodes in a series gate. The circuit should operate without changes on the TS-520, TS-820 and earlier models on the ICOM 701, and any rigs using similar blanker circuits, with minimal modifications to the rig. It has been used to good effect on my ICOM 701 for several months.

how it works

A block diagram of the circuit is shown in **fig. 1**. An MM5369 oscillator/divider integrated circuit, of the type used in many quartz clocks, together with a 3.579-MHz color TV crystal is used to generate an accurate 60-Hz square wave (reference 1). This signal is digitally divided by six by the CD4018 CMOS IC, resulting in a crystal-locked 10-Hz square wave. This 10-Hz signal is processed through a series of inverting CMOS Schmitt triggers (all contained in one 74C14 IC), the details of which are described below. The output of these stages is used to turn a transistor off and on. It is the collector of this transistor that is connected to the transceiver noise blanker, upon which it imposes a 10-Hz blanking pulse. The circuit diagram for the oscillator/divider stage is given in **fig. 2**.

delay and pulse-width circuits

To understand the workings of the 74C14 circuit, which forms the essence of this blanker, it is necessary to delve briefly into the arcane digital world of CMOS. Many Amateurs seem to have a fear of digital circuits and prefer to stick with good old analog tubes and transistors. There is really no good reason for this, as in many ways digital circuits are more predictable than analog ones. For those with no experience in CMOS, the *CMOS Cookbook* by Don Lancaster (reference 2) is a very good introduction. The operation of the 74C14 circuit is described in Chapter 4 of that book, and divide-by-six circuit in Chapter 6.

The easiest way to understand how the 74C14 circuit works is to look at what needs to be done to blank the Woodpecker. As described in the previous article, the object of this circuit is to provide a blanker control signal that is exactly synchronized with the Woodpecker. It should turn off the blanker gate only while the Woodpecker pulse is present, leaving the rest of the time between pulses for the desired signal to come through.

Thus we need a variable delay circuit to allow us to synchronize the blanking with the Woodpecker, and a means of varying the output pulse so that it blanks for no longer than necessary. It turns out that both these functions can be served by the same type of circuit, which Lancaster refers to as the "half monostable."

Consider the circuit in **fig. 3**. If a square wave is fed to the input, and the RC time constant is much shorter than the period of the square wave, the RC circuit differentiates; that is, it gives a positive spike when the input goes up, and a negative spike when the input goes down. As some CMOS circuits don't like negative input voltages, a diode is used to short out the negative spike. If this positive pulse is fed to a Schmitt trigger circuit (CMOS or otherwise) the output will be a narrow positive pulse, in synchronization with the rising edge of the input square wave.

If the RC time constant of the circuit is about the same as the period of the input square wave, however, the output is going to look like a sagged square wave; that is, the dc level does not decay very much

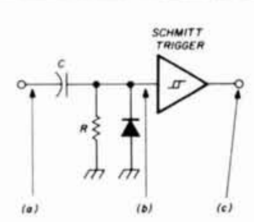

fig. 3. Half-monostable circuit.

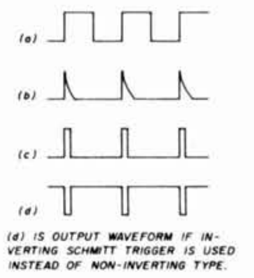

fig. 4. Waveforms for circuit in fig. 3.

before the square wave goes down again. Again, the diode cuts off the negative part of the wave. The waveforms are shown in **fig. 4**.

If this type of decaying square wave is fed to a Schmitt trigger, the output is a much broader pulse whose width is set by the point where the decaying voltage goes below the triggering level. As before, the beginning of the Schmitt output pulse is synchronized with the rising edge of the input square wave. Now if we make the resistor R a potentiometer, the RC time constant can be varied, and thus the width of the output pulse from the Schmitt trigger can be varied. It is most important to understand this, as the whole functioning of the synchronous blanker depends on this operation.

If you look up the specification sheet for the 74C14 in reference 1, you will notice that it contains six separate inverting Schmitt trigger circuits. Two points arise from this. The first is to point out the economy of using CMOS — only one IC is needed — and the second is that the output from an inverting Schmitt trigger is the inverse of an ordinary one. Thus, in a circuit such as **fig. 3**, the output of the inverting Schmitt trigger is positive all the time except for a brief drop to zero volts at the rising edge of the input square wave.

Okay, that should provide enough background to look at how the whole delay/pulse width circuit works. The circuit diagram for the delay and pulse width circuits is given in **fig. 5**.

The first stage in the delay/width circuit is a half-monostable (that is, as in **fig. 3**) using an inverting Schmitt trigger and a very short RC time constant (0.2 millisecond). The output from this stage is a brief negative-going spike synchronized with the rising edge of the input 10-Hz square wave.

The next stage is a half-monostable with a longer, variable RC time constant. The input to this is a square wave with a very high mark-to-space ratio; that is, almost all mark and no space. This waveform decays in the same way as described above, and the output from the inverting Schmitt trigger is a negative-going pulse whose width is variable from nearly zero to 0.1 second. The start of this pulse is also synchronized with the rising edge of the 10-Hz square wave input to stage 1.

Stage 3 is similar to stage 1: a half-monostable with a short, fixed RC time constant (0.2 millisecond). As before, the output is always positive except for a short drop to zero at the rising edge of the output from stage 2. Note now, however, that the output from stage 3 is synchronized not with the input to stage 1, but with the point where the decaying wave-

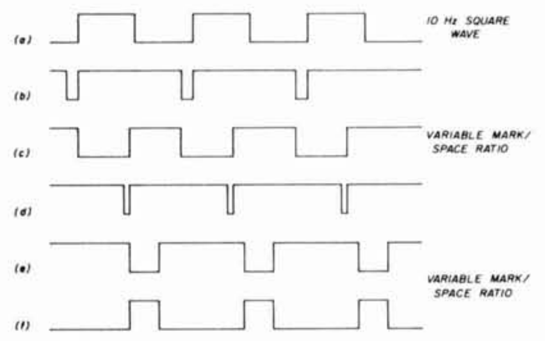

fig. 6. Waveforms for circuit in fig. 5.

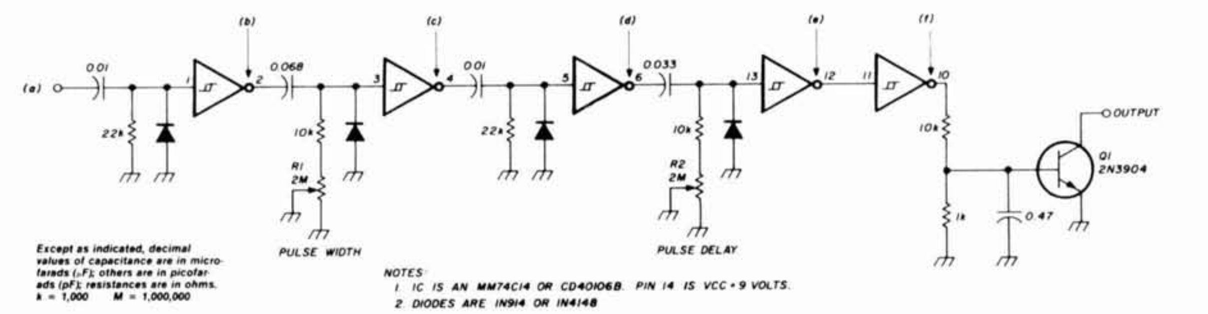

fig. 5. Delay and pulse-width circuits. Each device should be bypassed with a 0.01 μF ceramic capacitor with short leads.

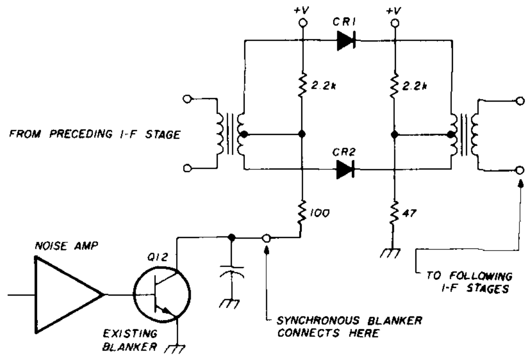

fig. 7. Simplified ICOM IC-701 noise blanker circuit (earlier models), showing point where synchronous blanker is connected.

form triggered the Schmitt in stage 2 to rise back to its positive value. This point is set by the variable RC constant, and thus by the 2-megohm potentiometer, R1.

In other words, at the output from stage 3, we have a brief negative-going spike whose phase with respect to the fixed, crystal-locked 10-Hz square wave input to stage 1 can be varied over a whole period (0.1 second) by R1.

Stage 4 is similar to stage 2; a half-monostable with a variable RC time constant. As with stage 2, the output is a negative-going pulse whose width is set by potentiometer R2. This is the pulse whose width determines how long the blanking goes on; that is, the blanking control pulse. The final Schmitt trigger circuit is merely used to invert the output pulse before it is applied to transistor Q1. The waveforms at various stages in the circuit (identified by letters in the circuit diagram) are shown in **fig. 6**.

Thus the final output from stage 4 is a pulse whose phase with respect to the fixed crystal-locked 10-Hz square wave can be varied by potentiometer R1, and whose width is set by potentiometer R2. This pulse is applied to the base of transistor Q1 through a current-limiting resistor. The pulse switches Q1 off and on, which allows it to alter the bias on the transceiver noise gate diodes, and hence blank the receiver.

IC701 blanker

A circuit showing the noise gate in the earlier model ICOM IC-701 is given in simplified form in **fig. 7**. The noise amplifier generates a control pulse that turns on transistor Q12 on the rf board. This action reverses the bias on diodes CR1 and CR2, and the signal is temporarily blocked from any further progress down the i-f chain. If Q1 of the synchronous blanker is connected in parallel with Q12 of the ICOM rf board, then the two transistors can operate independently of each other. In the ICOM it is a simple matter to tap into the rf board at the noise-blanker stage with a short length of coax to connect it to the external synchronous blanker.

Circuit diagrams of the Kenwood TS-820 and TS-520, and the Yaesu FT-75 suggest that a similar approach should work, and no doubt there are many others.

The newer ICOMs, however, use a 74LS123 to control the gate diode bias, and putting Q1 across the output of this device is not a good idea. In this case it would be necessary to use a circuit similar to that given in **fig. 8**, which involves a bit more tinkering and an extra 74LS01 IC.

adjustments

The main adjustment needed for the synchronous blanker is to ensure that the oscillator is accurately set. Trimmer capacitor C1 in **fig. 2** is used to tweak up the oscillator. The easiest way to do this is with a precision frequency counter. Pin 7 of the MM5369 in **fig. 2** is a test output at the crystal frequency. This should be set to 3.57954 MHz. Alternatively, if you have an accurately adjusted quartz clock using the MM5369 oscillator, it is possible to zero beat the two pin 7 outputs using a dual-beam oscilloscope.

The only other check necessary is to ensure that the waveforms look something like those shown in **fig. 6**.

how to use the circuit

The synchronous blanker is operated manually using the two potentiometers. When the Woodpecker makes its presence known, set the width control to about half way (corresponding to about 25 percent blanking of the incoming signal), and adjust the delay control until the interference is reduced as much as possible. Then readjust the width control to retain the blanking while attenuating the desired signal as little as possible. If all is well you should notice a significant reduction in the effect of the

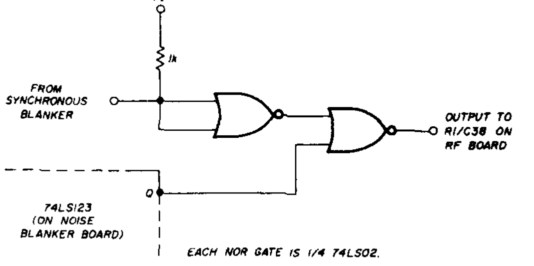

fig. 8. Modification to suit newer ICOM IC-701.

Woodpecker. If the oscillator is properly adjusted to give an accurate 10-Hz signal, the blanker should stay in sync with the Woodpecker for long periods without the need for readjustment.

A word of warning, however. While individual Woodpeckers stay synchronized for quite a long time, and even when returning after an absence will still be in sync, there are often occasions when more than one of the transmitters are operating in tandem. Under these circumstances, frequent readjustment is necessary and can be very tedious. Also, when the Woodpecker pulses are long and shaggy, it is necessary to blank so much of the incoming signal that intelligibility suffers severely. Under these circumstances, you have three choices: wait patiently, QSY, or go QRT!

Nevertheless, there are also times when the synchronous blanker is little short of miraculous. However, most of the time it falls somewhere between the two conditions, and provides a degree of respite.

conclusion

Most likely there are improvements that can be made to the circuit. The purpose of these articles is to draw attention to the possibilities of synchronous blanking, and see what others make of it.

In the next article in the series I discuss a further development of the circuit, which uses blanking in the audio stage. This obviously has the disadvantage that it does not overcome the AGC swamping effect of the Woodpecker; however, it can be fitted to any receiver, with or without noise blanker, with no need to get into the receiver circuitry.

references

1. *CMOS Databook*, National Semiconductor, 1978.
2. *CMOS Cookbook*, Don Lancaster, Howard Sams and Co., 1977.

appendix

The following was provided by Len Anderson, an assistant editor for *ham radio* and a contributor to our "Digital Techniques" series. Information is given on locating components, and some substitute devices are suggested. Also offered is an idea for using two chips to allow 10- or 16-Hz pulse switching. **Editor.**

component parts and alternates

All components should be available from many local and national sources. The MM5369 oscillator/divider is available from Radio Shack (part 276-1769). The MM5369 comes in three flavors, and an alternate is described below. The three versions are:

MM5369AA 60-Hz output (described by author)
MM5369EYR 50-Hz output (following divider must change)
MM5369EST 100-Hz output (following divider must change)

A suitable replacement could be the Intersil ICM7207 or ICM7207A, usually used with the Intersil ICM7208 counter IC to make a two-IC frequency counter. Source is Poly Paks or Newark Electronics.

The oscillator/divider could also be a part of a clock display using a National Semiconductor MM53107 clock chip. The MM53107AA has an auxiliary 60-Hz output; the MM53107FDU has an auxiliary 100-Hz output (divider change necessary). Caution: the MM53107 requires a 2.097152-MHz crystal and operates with a *2-6 volt* supply. The crystal can be obtained from Newark.

The 3.58-MHz crystal can be any NTSC TV color-oscillator crystal. The circuit will tune series or parallel resonance.

The CD4018B Johnson counter chip is a garden-variety CMOS device available from many sources. See **fig. A-1** for a suggested circuit to allow 10/16 Hz switching using two chips.

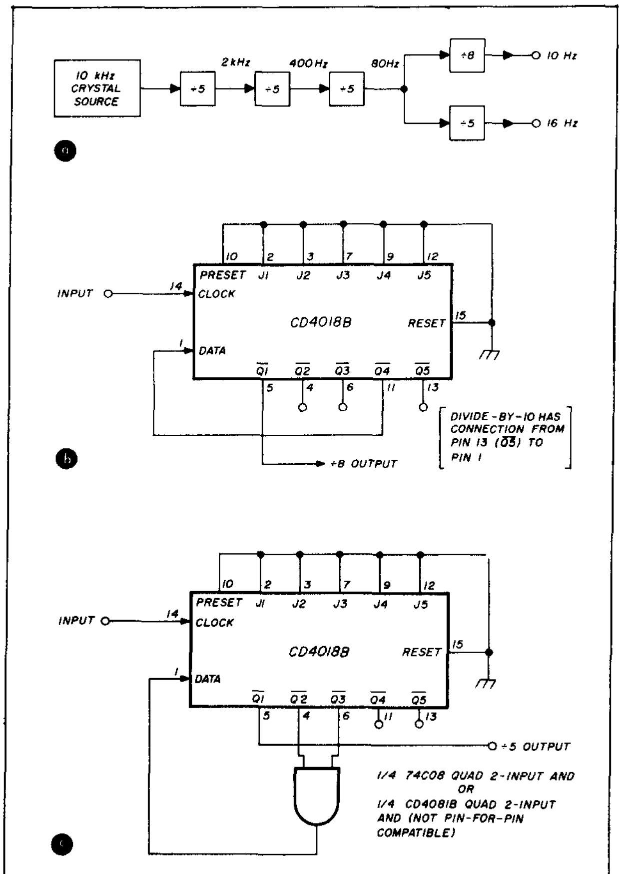

fig. A-1. Suggested circuit to enable 10- or 16-Hz pulse switching. The over-all counter chain is shown in *A*. The CD4018B is shown as a divide-by-eight and divide-by-five circuit in *B* and *C* respectively.

The hex Schmitt inverter can be either a 74C14 or a CD40106B. Both are CMOS and pin-for-pin compatible. Diodes in the delay and width circuits can be 1N914 or 1N4148. Both are small-signal, high-speed devices, readily obtainable. The 2N3904 can be replaced by several general-purpose NPNs; 2N3903, 2N4123, 2N4124, Motorola MPS2222 or MPS6514, or 2N2222.

alternative 10-Hz source

An existing frequency standard using a 1-MHz or 100-kHz crystal oscillator can be used as the timebase. Dividers must be used to get the 10-kHz repetition rate. North American power-line voltage can also be used as the 60-Hz standard. It will vary as much as ±0.1 percent only during changes in power flow (afternoons) but will otherwise be quite stable.

ham radio

an improved memory
for the versatile CW identifier

A better memory IC
— the 1702A EPROM —
in this follow-up
to a previous article

An earlier article in *ham radio* entitled "Versatile CW Identifier"[1] resulted in a tremendous reader response for which I am most grateful. One improvement that I've pursued is the use of a different memory IC in place of U3, an 82S126. Two reasons prompted me to seek a replacement for this chip. The first is the general unavailability of the 82S126 because of its age, and the second is the fact that it cannot be reprogrammed. By changing to a 1702A ultra-violet (UV) erasable programmable-only memory (EPROM), I can now more easily obtain the chips, reprogram them and, as an added benefit, have double the memory capacity.

the 1702A EPROM

The 1702A EPROM is a 2K-bit device arranged as 256 bits by 8 (in contrast to the 82S126, which is 256 by 4). This means that twice the number of messages can be programmed. The 1702A retains the information programmed into it when power is removed as does the 82S126, but by exposing the window on top of the 1702A to short-wavelength UV light for several minutes, all memory locations can be erased, and the chip is ready to be reprogrammed.

A programming device for the 1702A is fairly complex and expensive. Articles have been written describing its construction.[2,3] However, as with the 82S126, I am offering the benefits of a computer-controlled programmer that I've built (see **table 1** for more information). I ask only that the messages you request to be burned onto the 1702A follow the format outlined in reference 1 — with the only exception being that word spaces can be shortened to six bits instead of seven to accommodate a slightly longer message than would otherwise be possible.

table 1. Parts list for the improved memory circuit.

quantity	description
1	16-pin wirewrap socket
1	24-pin wirewrap socket
1	1702A ultraviolet erasable PROM
	wirewrap wire
1	vectorboard

Notes:
1. Those with 1702A chips may have them programmed by the author for $2.00.
2. Adapter kits, including all parts and programming of the 1702A, are available for $10.00.
3. Complete CW ID kits, with 1702A and adapter instead of 82S126, are available for $35.00.

All prices are postpaid. Allow 4-8 weeks for delivery. Write Michael J. DiJulio, 97 Woodside Road, Maplewood, New Jersey 07040. Please include a self-addressed stamped envelope for all inquiries not accompanied by an order.

modifications

I designed the new circuit so that anyone who has built the original unit can add the new chip easily. The modification consists of wiring a small adapter board, which consists of a 16- and 24-pin socket. In addition to this board, one part needs to be rearranged on the CW identifier board to provide the proper voltage interface to the 1702A. The 1702A requires both +5 volts and −9 volts (actually −7 to −9 volts will do).

A problem arises here, as the only voltages presently available on the board are +12 volts and +5 volts. Some type of dc-to-dc converter is needed. In the interest of simplicity, I used a little trick to obtain the two voltages without adding any parts. By isolating the +12 volt source ground from the system of the board, a separate +5 volt and −7 to −9 volt (depending on the +12 volt supply) referenced to the board ground can be made available (see **fig. 1**).

By Michael J. DiJulio, WB2BWJ, 97 Woodside Road, Maplewood, New Jersey 07040

To accomplish this modification, unsolder the 82-ohm 1W resistor (*R5*) and insert it, on end, into the holes originally used for the +12, −12 volt supply input (see **fig. 2**).

Cut a 4-inch (10-cm) piece of wirewrap wire and insert it into the hole originally occupied by *R5* that still connects to one end of *R5*. This wire is now the −7 to −9 volt supply line, which will go to the plug-in board. Attach a length of black hookup wire to this point by tack-soldering it to the island under the board. This is the −12 volt supply line (ground of the host vehicle, radio, or power supply).

Connect a length of red hookup wire to the other hole formerly occupied by *R5*, which is still connected to the cathode of *CR1*. This is the +12 volt connection that goes to *S1*. The important thing to remember from now on is that the −12 volt supply line (+12 volt ground return) must

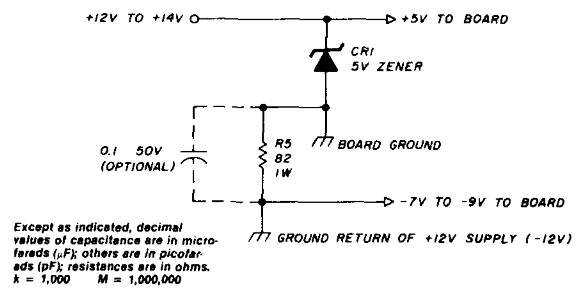

fig. 1. Power-supply modifications to provide −7 to −9 volts to the adapter board.

never be connected to the ground on the PC board! To do so will short out *R5*, disabling the −7 to −9 volt supply and almost certainly destroying *CR1*.

the ground system

You may ask how proper connection from the audio and keyer output jacks to the radio can be made with this seemingly haphazard ground system. Remember that the isolation between the two grounds is just an 82-ohm resistor, which, for the most part, will not alter the level of audio to the radio. However, for the purist a 0.1-microfarad capacitor may be shunted across *R5* to provide a more complete return path for the audio. With regard to the keyer output, in the original design a key closure looked like a 5-volt source in series with the same 470-ohm resistor; now it will look like a 12-volt source in series with the same 470-ohm resistor. The extra voltage should make no

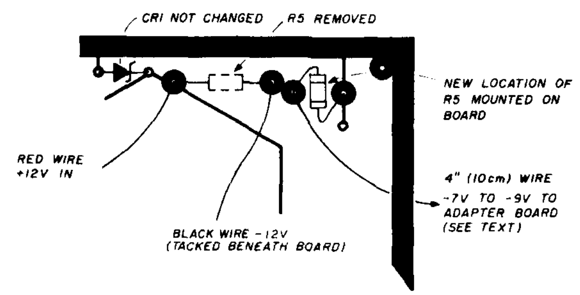

fig. 2. Details of the modifications made to the CW Identifier board to accommodate the new IC.

difference to the grid-block keying circuit of the transmitter.

As far as activating the circuit is concerned, *S1* can be replaced with a switched source of +12 volts such as the B+ line of the transmitter strip of the radio. Or if all that's available is the PTT line to ground, then replace *S1* with a piece of wire and use the PTT line to switch the −12 volt connection (the black wire previously mentioned). To summarize, no feature or performance is lost by modifying the power supply. However, be sure to isolate *all* ground board connections, such as the perimeter trace around the board, from the radio's ground system.

board construction

I toyed with the idea of laying out a PC board for the plug-in adapter board but felt that it was more trouble than it was worth. The simplest method for building the board is to obtain a 1.2 x 2.0-inch (30.5 by 50.8 mm) piece of perfboard with 0.1-inch (2.5-mm) hole spacing. Into this board, insert the 16- and 24-pin wire wrap sockets (**fig. 3**).

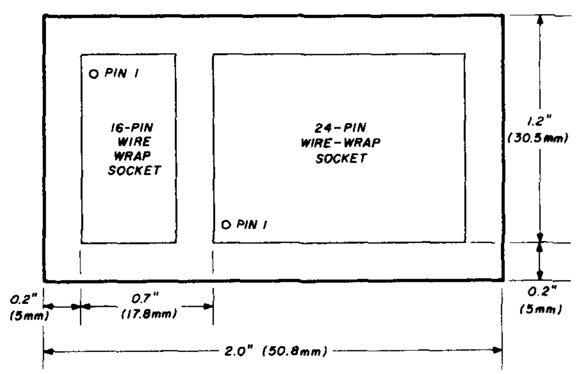

fig. 3. Top view of the plug-in adapter board showing arrangement for the 16- and 24-pin wirewrap sockets. Material is perfboard with 0.1-inch (2.5-mm) hole spacing.

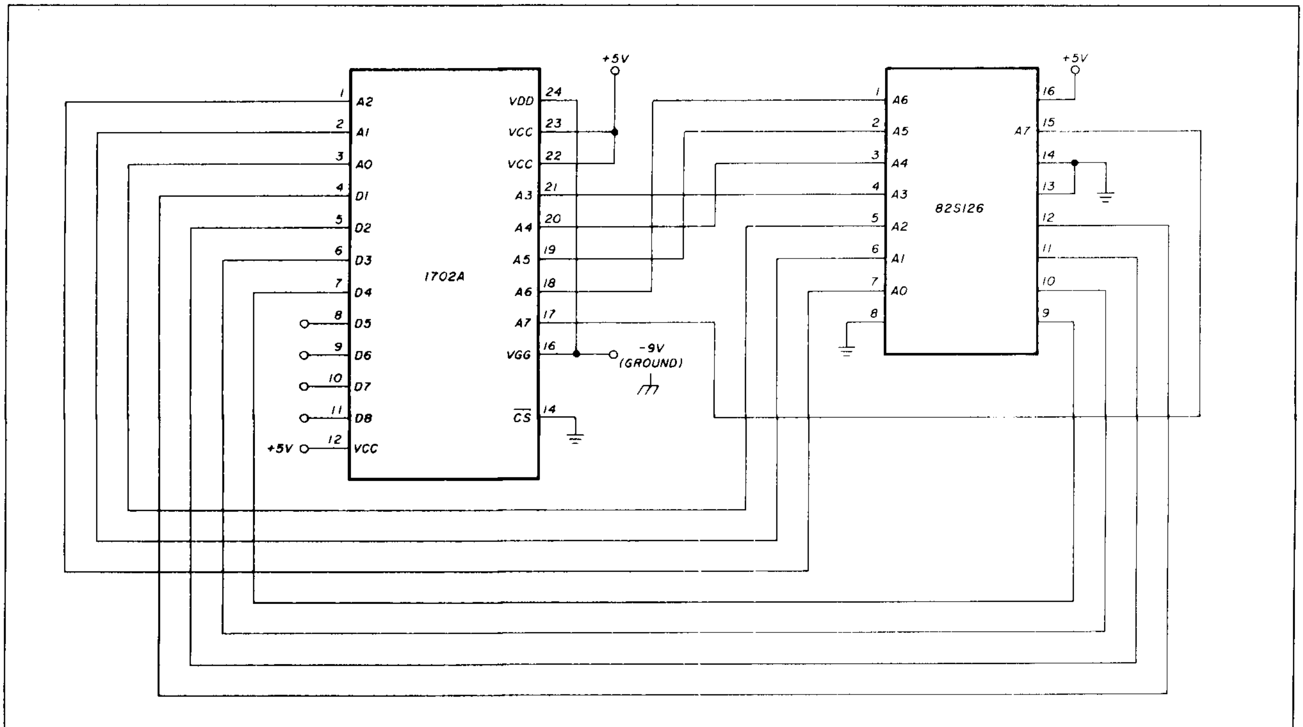

fig. 4. Schematic diagram of the adapter board for the CW identifier.

Using wire wrap wire and tools, connect the pins as indicated in **table 2**. (See also the schematic, **fig. 4**.)

Don't forget to connect the 4-inch (10-cm) piece of wire-wrap wire previously connected to *R5*. Instead of wire wrapping, the wire may be wrapped once around the pins and soldered to the pins close to the socket. Snip off the excess pin length (that is, just above the wire connection) on all pins of the 24-pin socket except for pins 8, 9, 10, and 11, which should be left at a length of 1/4 inch (6.4 mm).

Cut a small piece of thin cardboard measuring 1 by 1/4 inches (25.4 by 6 mm). Place this piece of cardboard on top of *Q1* and *Q2* to insulate the transistors from the adapter board. Now insert the 16-pin socket firmly into the socket that formerly held *U3* (82S126). Make sure pin 1 is lined up properly. Either the 82S126 or the 1702A may be inserted into this adapter. The original pads used for message selection are still used, but now pins 8, 9, 10, and 11 of the 1702A permit the selection of four more messages when the 1702A is used.

I believe this modification is a simple and worthwhile one, which will greatly improve the utility of the identifier.

table 2. Wirewrap connections between the 16- and 24-pin devices.

16 pin socket	→	24 pin socket
1		18
2		19
3		20
4		21
5		3
6		2
7		1
9		7
10		6
11		5
12		4
14		14
15		17
16		12, 22, 23

Note: 16 and 24 to − 9 volts on main board.

references

1. Michael J. DiJulio, WB2BWJ, "Versatile CW Identifier," *ham radio*, October, 1980, pages 22-25.

2. Dan Vincent, "Low-Cost EPROM Programmer, Part I," *Popular Electronics*, February, 1978, pages 41-45.

3. Dan Vincent, "Low-Cost EPROM Programmer, Part II," *Popular Electronics*, March, 1978, pages 55-58.

ham radio

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This fine receiver
has been the target
of many design improvements:
here's another

Drake R-4C receiver
improved power supply

In keeping with our policy to provide the most recent information on updating equipment, we present this article on improvements to the popular Drake R-4C communications receiver. As author Klinman points out, the perfect receiver has yet to put in an appearance. The R-4C by Drake with its many features comes close to the perfect receiver. These modifications to the R-4C are easy to make, result in a significant operational improvement, and use a minimum of mechanical modifications to preserve resale value of the radio.
Editor

The perfect communications receiver has yet to be produced. The Drake R-4C approaches this ideal. With certain modifications it can be made into a real "performance" receiver. Among these are the addition of selective first i-f filters, improved product detector and audio modifications as described by Sherwood,[1] third mixer redesign with solid-state tube replacement,* an audio lowpass filter by Sartori,[2,3,4] and agc modification by Klinman.[5] I recommend that those using the Drake R-4B/C obtain the excellent summaries of updates to these receivers available from Sherwood Engineering and Sartori Associates.†

*While the "Solid Tube," a product of Sartori Associates, used as replacement for the 6EJ7 mixers in the R-4C does effectively eliminate the severe noise generated by the vacuum tubes in these circuits, it does noticeably reduce large-signal-handling capacity of the receiver.

†Sherwood Engineering, Incorporated, 1268 South Ogden Street, Denver, Colorado 80210; Sartori Associates, P.O. Box 2085, Richardson, Texas 75080.

By Richard Klinman, W3RJ, RD 1, Flint Hill Road, Coopersburg, Pennsylvania 18036

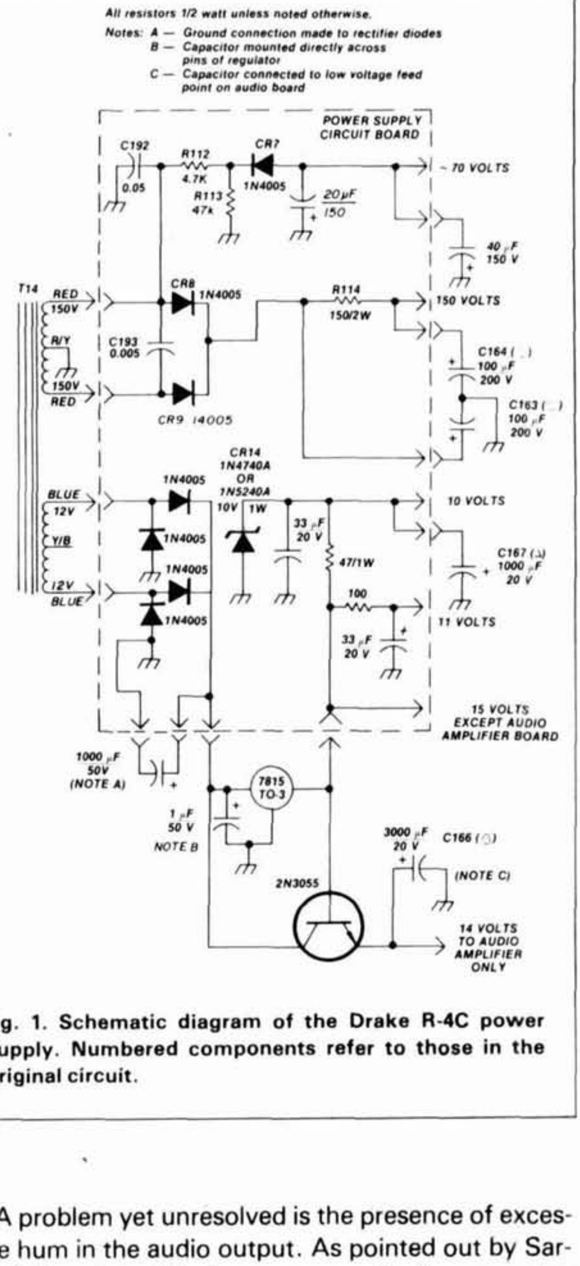
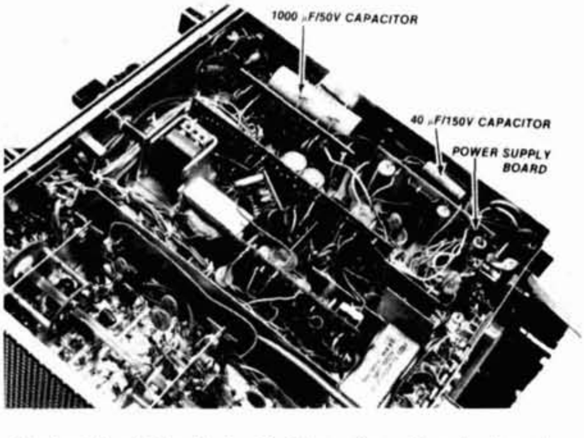


fig. 1. Schematic diagram of the Drake R-4C power supply. Numbered components refer to those in the original circuit.



Underside of the Drake R-4C receiver chassis showing revised power-supply board and component layout.

regulator, but the monolithic circuit is difficult to stabilize and it can still require considerable peak current when driving a low-impedance load.* An advantage of the monolithic audio power amplifier is that it provides a significant amount — on the order of 30 dB — of power-supply ripple rejection.

As described here, it's a relatively simple matter to retain the original R-4C audio amplifier and upgrade the power supply.

revised power supply

Fig. 1 is the schematic diagram. A full-wave bridge rectifier with single-stage capacitor-input filter produces 25-30 volts, which is sufficient to power the 7815, a 15-volt monolithic voltage regulator. While increasing the average power dissipated by the power transformer, T14, no additional temperature rise of the transformer is noticeable. Because this supply voltage exceeds the voltage rating of the

A problem yet unresolved is the presence of excessive hum in the audio output. As pointed out by Sartori,[3] the original Drake power supply also generates significant heat under the R-4C chassis. In that article, Sartori described modifications to the power supply, but the suggested circuit produces marginal voltage for proper operation of the recommended monolithic voltage regulator. In addition, the circuit yields a regulated low voltage 2 volts lower than the original 14-15 volts.

Complete replacement of the R-4C audio amplifier with a monolithic audio power amplifier, as suggested by Sherwood,[6] will reduce the audio-amplifier average current drain on the low-voltage supply and

Rear apron of the Drake R-4C receiver showing the voltage regulator and pass transistor, heat sinks, and homebrew mounting plate.

*The circuit may be stabilized by using Sherwood's output stabilization network, which is part of their audio-amplifier kit. **Editor**

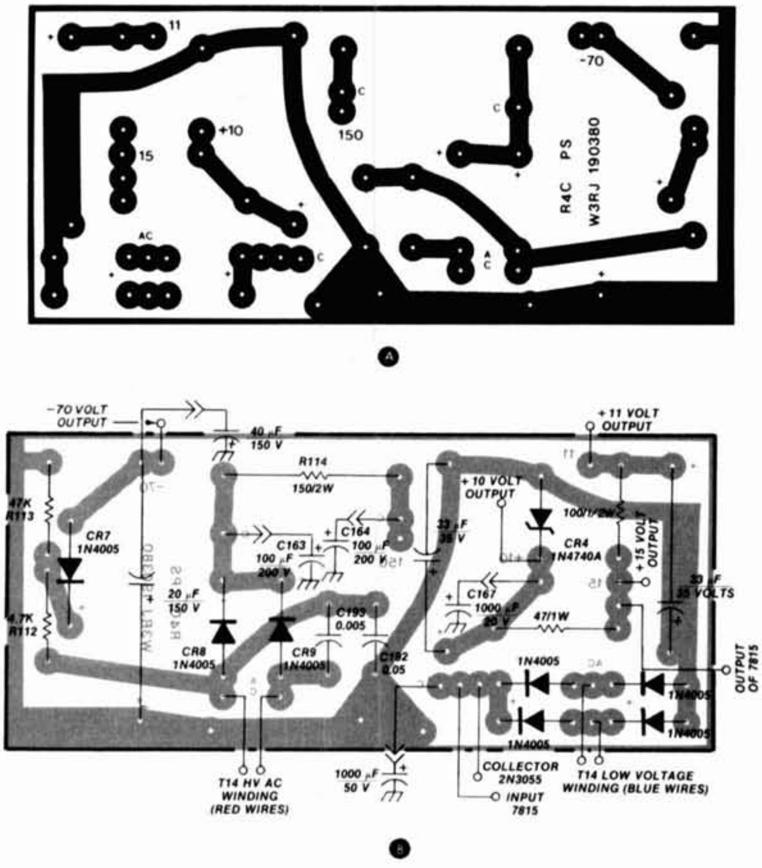


fig. 2. Circuit-board details. Full-size PC-board mask is shown in (*A*). Illustration (*B*) shows component layout.

existing filter capacitors, C166 and C167, an additional 1000 µF 50-volt electrolytic filter capacitor is required. The output of the TO-3-cased 7815 (or equivalent) 1-amp 15-volt regulator powers all 15-volt circuits *except the audio output stage*.

Power for the audio amplifier cannot be taken directly from the 7815 because of severe instability caused by the large inductive load of the audio output stage. A common base 2N3055* pass transistor provides the required isolation in addition to reducing the thermal load on the 7815 voltage regulator. One of the unused sections of the R-4C filter capacitor, C166, is used to further reduce the impedance of the 14-volt supply to the audio amplifier. A series resistor provides 11 volts for the PTO, and a zener diode provides 10 volts for the BFO and HFO circuits.

The remainder of the power supply is similar to the original circuit in the R-4C. An exception is additional filtering of the −70 volt supply to eliminate the last trace of hum in the receiver audio.

hardware

A board containing all components except the pass transistor, voltage regulator and 1-µF bypass capacitor, and the 1000-µF, 50-volt and 40-µF 150-volt filter capacitors replaces the original R-4C power supply board. The board mask and component placement are shown in **fig. 2**. Solder lugs, bent 90 degrees in the center and soldered to the board at the indicated locations, serve as mounting feet similar to those on the original board. Noteworthy details of construction are:

1. The low-voltage filter capacitor (a small size 1000-µF, 50-volt electrolytic) is mounted between the power supply or audio circuit board and the side of the chassis. It is secured to the chassis side wall with a plastic stick-on cable tie anchor. The ground lead

*Any NPN power transistor with V_{ceo} of at least 40 volts and collector current rating of 1 ampere will work in this circuit. (The 2N3055 is available for less than a dollar from several mail-order parts houses.)

of this capacitor must be returned to the ground foil of the power-supply board. A chassis ground to the filter capacitor cans, C163 through C167, while mechanically convenient, will lead to audio hum.

2. The 40-μF, 150-volt miniature electrolytic capacitor filtering the −70 volt supply is mounted between the power-supply board and the side of the chassis. It is soldered from the −70 volt output pad to the ground on the foil side of the board and is supported by its axial leads.

3. The centertap of the low-voltage winding (yellow/blue) of the power transformer, T14, is disconnected from ground, covered with heat-shrink tubing at the end to avoid short circuits to ground, and tucked out of the way along the chassis.

4. Connection between C166 and the 14-volt supply to the audio board is made to the low voltage supply solder lug directly on the audio board. The photograph shows the new power-supply board and component layout under the chassis.

The 7815 monolithic voltage regulator and 2N3055 pass transistor are mounted to the rear apron of the Drake R-4C on a small homebrew 2-1/2 × 3-3/4 × 1/2-inch (63.5 × 95 × 12.5-mm) open-bottom box.* This shallow box is fastened to the rear of the Drake R-4C with a pair of 1/2-inch (12.5-mm) threaded standoffs (photo). In this way, only two small, inconspicuous holes need to be drilled in the Drake R-4C. Wires are cabled and routed through a grommet inserted into one of the slots in the back apron and through the power transformer grommet, to the underside of the chassis. Small, finned heatsinks of the Walefield 680 type cool both voltage regulator and pass transistor.

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*This miniature open-bottom chassis can be easily fabricated from 1/16-inch (1.5-mm) aluminum sheet. Alternatively, the regulator and transistor may be mounted on a flat rectangular 2-1/2 × 3-3/4-inch (63.5 × 95-mm) plate. The 1/2-inch (12.5-mm) lip covers the electrical connections to the regulator circuit.

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In part three we discussed some of the active electrical devices and their uses. Active devices alter the voltages or currents applied to them in some way. Usually, the resulting currents and voltages are nonlinear, or distorted in waveshape to a greater or lesser extent. The active devices we covered were vacuum tubes and solid-state diodes. This month we will discuss shunt regulation, junction transistors, series regulation, special semiconductor diodes, FETs, SCRs, and triacs.

simple shunt regulators

To produce low-impedance (constant voltage) power supplies, it is necessary to add some form of voltage regulation. If a regulated 12-volt supply is required, it may be necessary to use something like a 20-volt unregulated dc supply and add regulation circuits between the unregulated voltage and the load.

A simple shunt regulation circuit uses a zener diode. This is a diode which has heavy doping of its semiconductor materials. In its forward direction it acts much like a normal semiconductor diode. As the inverse voltage is increased across a zener diode a point will be reached (the zener, or avalanche, voltage) where the diode suddenly allows current to flow in the reverse or inverse direction. Such an inverse avalanche current in a normal diode would destroy the crystalline form of the semiconductor materials and ruin the diode. But the zener will withstand whatever current it is manufactured to handle. A diagram of a zener-diode-regulated 12-volt power supply is shown in **fig. 1**. If the zener is manufactured to pass 1 amp of current safely, series resistor R_S should have the necessary value to pass not only 1 amp, but at the same time to produce a voltage drop across it of 20 volts minus 12 volts, or 8 volts. Therefore the

By Robert L. Shrader, W6BNB, 11911 Barnett Valley Road, Sebastopol, California 95472

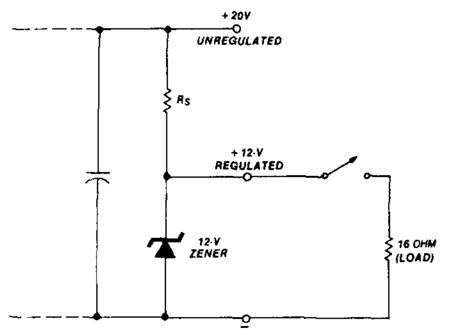

fig. 1. Using a zener diode to provide a regulated 12-volt dc output.

value of R_S is computed to be $R = E/I$, or 8/1, or 8 ohms. Its wattage rating would have to be at least $P = EI$, or 8(1), or 8 watts. Any 8-ohm resistor with a power rating over 8 watts would be satisfactory.

If a 16-ohm load is now connected across the zener's 12-volt voltage-drop, the current for the load would be $I = E/R$, or 12/16, or 0.75 amp, leaving the zener with 0.25 amp flowing through it. Since the zener still has 12 volts across it with 0.25 amp flowing through it, while the load is drawing 0.75 amp, the current through R_S must still be the same as with no load. The load can vary up and down from its 16-ohm value quite a bit without producing any change in the voltage across it. However, if the load increased (lowered resistance) to an effective 10 ohms, the zener would lose its control. Now the total resistance across the 20 volts would be 8 + 10 ohms, or 18 ohms. With 18 ohms and 20 volts the current would be $I = E/R$, or 20/18, or 1.11 amps. With 1.11 amps through the 10-ohm load, the voltage-drop across the load would be $E = IR$, or 1.11(10), or 11.1 volts. This is below the 12-volt zener voltage of the diode and it would no longer be carrying current and could no longer hold the output voltage at 12 volts. (We have disregarded the fact that the 20 volts would reduce under load.) If R_S were made 5 ohms instead of 8 ohms, the zener would regulate the 10-ohm load for reasonable variations, but if the load were disconnected or lightened excessively the current through R_S and the zener would increase and possibly burn out the diode. For many low current circuits, zener diodes do a good job of voltage regulation.

A similar type of regulating circuit is used with higher-voltage vacuum-diode-rectifier power supplies. Instead of zener diodes, they use gaseous voltage regulator (VR) tubes. These tubes have two metal elements in them, and are filled with a gas such as neon. At about 90 volts neon ionizes (and glows) and allows current to flow through it, but at a voltage-drop of approximately 75 volts across the tube. Over a current range of 5 to 35 mA a VR 75 tube will hold its voltage-drop at close to 75 volts. Thus, VR tubes can be used to regulate loads that vary between 5 and 35 mA. VR tubes come in 75-, 90-, 105-, and 150-volt values. To regulate a 300-volt output, two VR-150 tubes in series could be used. The unregulated supply must provide a no-load voltage of about 360 volts to produce the required initial ionization to start current flowing in the VR tubes in series to regulate 300 volts. (Similarly, two 12-volt zeners in series can be used to regulate to 24 volts.)

junction transistors

One of the transistors in use today is the bipolar junction transistor. We will call it a BJT to differentiate it from a field-effect transistor, or FET. A BJT is made up of three pieces of P and N semiconductor, somewhat as shown in **fig. 2**. In this particular NPN device there is a single thin P-material section between two pieces of N-material. Note the similarity of this triode (three-element) circuit to the VT (vacuum tube) triode circuit — but also note the differences.

The collector (C) circuit supply, V_{CC}, tries to forward-bias the emitter-base (EB) junction, since it has its negative terminal to the lower N element. But the upper PN junction is reverse biased by this same V_{CC}. As a result, no collector current (I_C) can flow. In this circuit the lower PN junction is being forward biased by the base power supply, V_{BB}. With the EB junction forward biased there is a current of electrons spreading out in the P material trying to find a path through the relatively poorly conducting semiconductor material to get from the emitter to the base connection. In spreading out looking for positive holes to use as stepping stones through the P material, some of these electrons move into the barrier area of the upper PN junction. Electrons in the upper barrier area cancel the barrier effect there, thus allowing current to flow from emitter to collector. The more emitter-base current that flows the more the base-collector

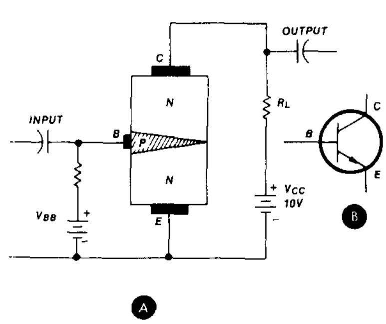

fig. 2. (a) Basic bipolar junction transistor in an amplifier circuit. (b) Symbol of an NPN type BJT.

barrier is cancelled and the greater the emitter-collector current can be. If the EB current decreases, so does the EC current. In a transistor of this type it may be found that a small EB current change in the input or base circuit can produce a 100 times greater EC current change in the output or collector circuit. This represents an amplification of current variation (h_{FE}), which is somewhat similar to the voltage amplification (mu) that occurs in a VT triode. We say the VT is voltage actuated, because voltage variations in the grid circuit control output circuit current (and power). In a BJT it is current changes in the input base circuit which control output current, so it is said to be current actuated.

Since the EB and the EC currents are both in the same direction through the emitter, the emitter current is always equal to the sum of the base and the collector currents, or $I_E = I_B + I_C$.

When amplifying ac signal currents, normally shown by lower case letters, the ratio of i_c/i_b is known as beta (β). On the other hand, the ratio of i_c/i_e is the alpha (α) gain of the transistor. At higher frequencies the alpha value decreases. The frequency at which it decreases to 0.707 (also known as 3 dB decrease) of its value at 1000 Hz is known as the alpha cut-off frequency (f_{hFB}).

Rather than using a separate battery or supply to produce the forward biasing current of the base circuit, it is simpler to run a relatively high value of resistance (100 to 250 kilohms) from $+V_{CC}$ directly to the base. This is shown as R_1 in **fig. 3a**. The value of this resistor is that required to reduce the forward biasing base current to about 1/100 of the normal I_C at which the transistor is to operate. For example if I_C is to be 0.005 amp, then I_B should be $0.005 \times 1/100$, or $0.00005\ A$, and $R_1 = E/I$, or $10/0.00005$, or *200,000 ohms*. (Consider the EB junction to be essentially zero ohms.)

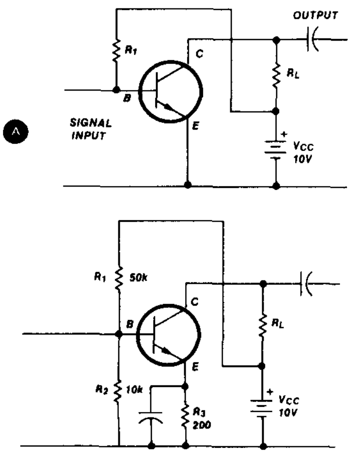

fig. 3. (a) Simple forward biasing method for a BJT amplifier. (b) Improved biasing circuit for a BJT.

While this simple biasing circuit will work, a far more satisfactory circuit, and the one generally used, is shown in **fig. 3b**. One of the difficulties with transistors is that they are made of semiconducting materials which always have a positive temperature coefficient (+TC). As the BJT operates, its I_C tends to heat it and it becomes a better conductor, so more current flows and it heats still more. If this cycle continues it can result in thermal runaway and the junctions of the BJT may melt. In the circuit of **fig. 3b**, the voltage divider biasing provided by R_1 and R_2 gives a more stable biasing. When I_C flows up through R_3 it produces a voltage-drop across this resistor. The voltage developed is negative at the bottom and positive at the top of R_3. When considered in series with R_2 and the base circuit, this voltage-drop will be seen to be a reverse bias for the base, which tends to reduce I_B and therefore I_C. So, with this network of resistors, when I_C begins to increase due to thermal effects, the reverse bias increases and the rise in I_C is held in check. However if the BJT is allowed to heat too much it may still self-destruct, and it can do this unbelievably fast! The ratio of R_1 to R_2 is usually about 5/1, perhaps being 50,000 and 10,000 ohms as the working values. R_3 is usually in the neighborhood of 200 to 500 ohms in low-power circuits (less in high-power circuits).

The reverse bias developed across R_3 will also occur at whatever the signal voltage or current rate is, resulting in a lessening of the effective input signal and a lessening of the output signal from the whole "stage" (BJT and all the components). The capacitor C_1 acts as a filter capacitor and tends to hold the voltage-drop constant as far as the rapid signal ac frequencies are concerned, but has no effect on slow thermal variations. Its value will vary from 0.01 for rf circuits to several microfarads for audio-frequency circuits.

Thus far, we have discussed just one type of transistor, the NPN bipolar junction transistor. BJTs can also be made with a P-material emitter, an N-material base, and a P-material collector. They are then known as PNP BJTs. They operate just like NPNs, except that the collector bias voltage, V_{CC}, is negative toward the collector, rather than being positive as in the NPNs discussed so far. It is not unusual to see both NPN and PNP transistors used in the same circuit to take advantage of the complementary action of the two devices. The arrow on the symbol of a PNP transistor is reversed from that of an NPN.

Most transistors are low-power devices. When higher power is required, BJTs must be made larger and must have some means of keeping the base-collector junction from getting too hot. This is accom-

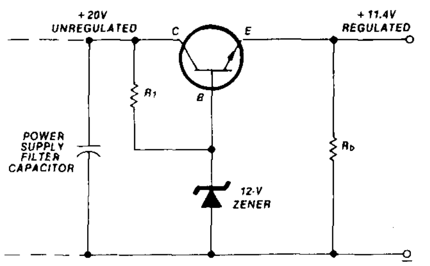

fig. 4. Series pass transistor voltage regulator circuit.

plished by using metal *heatsinks* on the transistors. Sometimes the heatsinks are built into the transistors and sometimes they must be added over the top of the body of the transistor. The bodies of power transistors are made with the collector element nearest the surface of the device to allow optimum cooling of the CB junctions.

series regulation

A BJT can be used to regulate the voltage output of an unregulated power supply, as shown in **fig. 4**. With R_b connected, there is a small I_C flowing in the transistor. A silicon EB junction will have a 0.6-volt barrier voltage-drop, making the output voltage equal to the 12 volts of the zener minus the 0.6 volt of the EB barrier voltage-drop for an 11.4-volt output. If a load is added, the I_C will increase and the voltage-drop across EB should increase slightly. This appears to the BJT as a greater forward bias and it passes more current for the load, and the regulated voltage approaches 11.4 volts again. However, such a regulation is never complete. There is always some change in output voltage when the load changes, which can be called an error voltage. In this simple series regulator the error voltage may be a fraction of a volt. If the error voltage can be amplified and fed to the base of the series regulator "pass" transistor, the output voltage variation will be better corrected and the power supply will be said to be better regulated.

In the series transistor regulator circuit of **fig. 5**, Q_1 is the pass transistor and Q_2 is the error voltage amplifier. R_B forward biases Q_1 and also acts as the collector load resistor for Q_2. R_Z is the zener resistor to hold it in conduction and provide a regulated reference voltage. The position of the arm on the potentiometer will determine the regulated output voltage within a limited range. If a load is suddenly applied to the power supply the output voltage of the supply tends to become lower. The base of Q_2 feels the reduced forward bias, which reduces its I_C. With less Q_2 collector current flowing through R_B there is less voltage-drop across it. The lessened R_B voltage-drop appears to the Q_1 base as a greater forward bias, and it passes greater I_C for the load. This re-establishes essentially the same voltage output as with no load. The regulation is perhaps 50 to 100 times better with this circuit than with the simpler circuit of **fig. 4**. The capacitor shown dashed is used as a filter capacitor to prevent sudden changes of voltage from being fed to the Q_2 base, which might cause excursions of the output voltage. Regulator circuits of this type are quite popular in transistorized Amateur Radio equipment. For better regulation under varying temperature conditions, added components are required which can complicate the circuits considerably.

Pass transistor voltage regulators can be made up into small *integrated circuit* (IC) packages about 3/8 × 3/8 × 1/8 inch (10 × 10 × 4 mm) that can regulate up to 1 ampere or more. They have three leads, one for the input unregulated positive voltage, one for the output regulated positive voltage, and one for the negative or ground lead. Additional 1-μF capacitors must be connected between input and ground and output and ground leads to prevent external ac fields from damaging the internal IC circuits. Regulator circuits made up of separate transistors, zeners, resistors, and capacitors are known as discrete circuits, indicating that they are constructed from separate, or discrete, parts.

Series regulation circuits for VT power supplies are similar to solid-state circuits. They require separate filament windings on the power transformer for the triode or tetrode pass tube and for any error amplifier. VR tubes are used as the reference voltage devices.

special semiconductor diodes

Besides the semiconductor diodes used as rectifiers and zener regulators, there are several other commonly used solid-state diodes.

The *tunnel diode* is a specially doped diode that has an unusual voltage-current curve. From zero

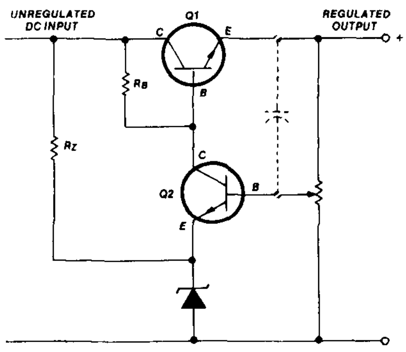

fig. 5. Pass transistor regulator capable of very good regulation.

volts forward bias to a fraction of a volt the current increases as voltage increases. Then, with a little more voltage increase the current begins to decrease. With still greater voltage increase, a point is reached where the current again begins to increase with an increase in forward biasing voltage. If operated in the downward sloping part of its curve it is said to be operating in a *negative resistance* condition, since current normally increases in a device when the voltage across it increases. Negative resistance can be useful in developing oscillator (ac generator) circuits and for special amplifiers with such diodes. The power output of tunnel diodes is quite small but they can operate at or generate extremely high frequency ac.

A *voltage-variable capacitor diode* (varactor, varicap) is used as a variable capacitor. You may remember that when reverse biased, a standard semiconductor diode develops a barrier area at its junction. The width of the barrier area depends on the value of the reverse bias voltage. Since a PN diode in a reverse biased condition is two conducting substances separated by a nonconductor (the barrier area), increasing the reverse bias increases the barrier width and reduces the capacitance between the two P and N materials. Thus, by varying the dc *reverse* bias on a varactor the capacitance it exhibits can be controlled. A varactor operated across an LC circuit can change the resonant frequency (that is, tune the LC circuit).

A *hot-carrier diode* (HCD) is one in which a semiconductor is grown against a piece of metal. Such a diode exhibits a rectifying action, but no barrier area is developed. As a result, the junction between semiconductor and metal can be instanteously made to pass or stop current flow. In the normal PN junction there are always a few misplaced electrons which act as "minority carriers." These must be swept out of the barrier area before the diode can turn on or turn off its current. This limits the speed of operation, or the frequency of the ac which the diode will rectify efficiently. Hot-carrier diodes are also known as *Schottky diodes*, and are useful up to super high frequencies.

A PIN *diode* has a P-material block separated from an N-material block by a thin intrinsic (I, or undoped) layer. These devices act as diodes up to a megahertz or so and then the intrinsic layer's slow action results in too much transit time delay. Above this frequency PIN diodes are useful because they have a high resistance when reversed biased, and a low resistance when forward biased. If inserted in a transmission line carrying high-frequency rf or microwave energy they can be made to effectively stop or allow the energy to pass down the line. We say the microwave energy transmission can be modulated (varied) by changing the PIN diode biasing voltages.

Point-contact diodes were the first semiconductor devices. The original ones were the crystal detectors of the early 1900s and are still available. They are somewhat similar to HCDs. They were improved during World War II and found use as detectors in radar sets. They are still used to detect microwave and other lower frequency rf signals. A point-contact diode consists of a fine wire cat-whisker touching a "sensitive" point on some form of metallic semiconductor. The original transistors were actually two separate point-contact diodes close together on a single piece of germanium or silicon.

A *light-emitting diode* (LED) radiates light when forward biased. All diode junctions when forward biased emit some frequency of radiation, usually in the invisible infra-red range. In most cases the radiation is shielded from the outside except at any exposed edge of the junction. With specially selected semiconductor materials it is possible to produce diodes that emit red, orange, yellow, green, or blue light waves. The diodes are manufactured to provide a maximum junction area to be visible, so that the light wave radiations will be emitted efficiently. LEDs also act as rectifying diodes, but have slightly higher barrier voltages than normal silicon diodes.

field-effect transistors

A type of transistor equally as important as the BJT is the field-effect transistor (FET). There are several types. One is the junction field-effect transistor (JFET). Another is the metal-oxide semiconductor field-effect transistor (MOSFET), of which there are several varieties.

The device shown in **fig. 6a** is a simplified N-channel JFET. It is called an N-channel device because there is an N-material channel between two P-material areas which are connected together electrically. At one end of the channel is a contact called the *source* (S). At the other end is the *drain* (D) contact. Any reverse bias in the *gate* (G) circuit (from G to $-V_{GG}$ to S) widens both barrier areas, which effectively pinches the channel narrower. The narrower the channel the more resistance it has and the less source-drain current that will flow with a given amount of drain circuit voltage ($+V_{DD}$). If sufficient negative bias is applied to the gate, the drain current (I_D) will be pinched off completely. If the gate is made positive, gate current will flow from gate to source. If you analyze this biasing action you will see that it is essentially the same as occurs in a vacuum tube as far as I_D and I_P actions are concerned. Thus, *a JFET is a voltage operated transistor, whereas NPN and PNP BJTs are current operated.* Within limits, JFETs can be used in the same types of low-power circuits in which VTs are used, and vice-versa.

The JFET described is an N-channel type. If a JFET is manufactured with a P-channel between the two N-material areas or substrates, it is called a P-channel JFET. Basically the only difference in operation between N- and P-channel JFETs is the polarity of the V_{DD} and V_{GG} used with them. The symbol for an N-channel JFET is shown in **fig. 6b**, and for a P-channel JFET in **fig. 6c**.

An *insulated-gate* FET (IGFET, MOSFET, CMOS, COS/MOS, VMOS, etc.) has a thin N-channel grown on the surface of a P-material substrate, **fig. 7**. A metal contact is cemented to the N-channel with a thin layer of insulating material. If this metal gate (G) is made more negative, it attracts the positive holes from the P-material, which pinches the channel thinner, raising the N-channel resistance and reducing the I_D value. Variations of the gate voltage produce variations of the drain current, again very much like the operation of a VT, except that the gate may be driven positive and not draw any current from the channel because of the insulation layer. This is known as a depletion type MOSFET. There are also enhancement MOSFETs, which operate somewhat similarly except for the bias voltages used. In general, depletion MOSFETs use either a reverse direction bias similar to a VT, or zero bias. An enhancement MOSFET uses a forward direction bias voltage to which is added the input signal voltage to be amplified.

There are also dual-gate MOSFETs in which either or both gates can help to control the drain current, **fig. 7c**. These are used in *mixing circuits*, discussed in later articles.

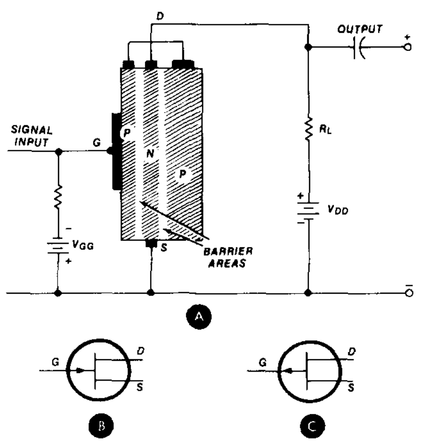

fig. 6. (a) Basic junction field-effect transistor in an amplifier circuit. (b) Symbol for an N-channel JFET. (c) Symbol for a P-channel JFET.

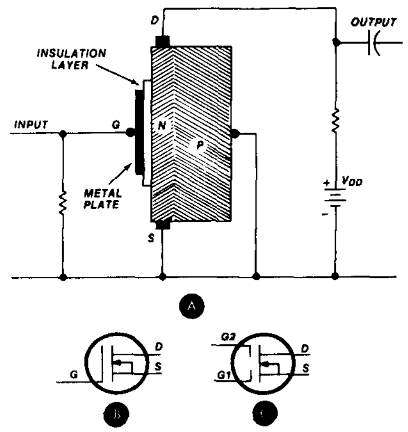

fig. 7. (a) Basic MOSFET in an amplifier circuit. (b) Symbol for an N-channel MOSFET. (c) Symbol for a dual-gate N-channel MOSFET. The P-type base substrate may or may not be connected to source or ground.

One of the newest high-frequency, high-power transistors is the VMOS, which is a special form of the MOS transistors. It gets its name from the V-shape channel used in it.

Note the similarities of the basic operations of the three elements of VTs, BJTs, and FETs:

VT		BJT		FET
Cathode (K)	=	Emitter (E)	=	Source (S)
Grid (G)	=	Base (B)	=	Gate (G)
Plate (P)	=	Collector (C)	=	Drain (D)

There is another type of transistor, called a *unijunction* transistor (UJT). It is a bar of N-material of about 10,000 ohms resistance, with a small emitter area grown on its side about half way down the bar. When the emitter bias is increased there will be a voltage at which the bar very suddenly becomes a good conductor. When the emitter bias is reduced there will be another value where the resistance of the bar very suddenly jumps back to the high resistance value. Unijunction transistors are not amplifying transistors, but can be used to trigger other circuits. They obtain their very fast change of state due to the negative resistance effect that occurs in them, which develops a form of regeneration. Regeneration in any device or circuit always tends to speed up the action time of circuits in which it is present.

SCRs and triacs

A *silicon-controlled rectifier* (SCR) is an NPNP, or four-layer semiconductor device. It has a cathode (K), an anode (A), and a gate (G), as indicated in the SCR circuit in **fig. 8**. In this circuit, with the switch open, regardless of which half of the ac is considered, at least one of the NPNP junctions will be re-

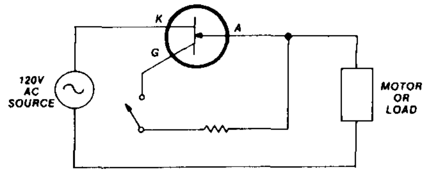

fig. 8. SCR as the control device in a motor circuit.

verse biased, preventing any current from flowing in the motor or load. When the switch is closed, one of the junctions that would normally be reverse biased is now forward biased and current can flow through the SCR, turning the motor. With an SCR, however, only current in one direction will be flowing through the load because of the one-way rectifying action.

A *triac* is another type of breakdown or *thyristor* NPNPN device. It acts like two SCRs back to back so that when the gate is activated it will pass current in both directions, terminal to terminal, through itself. The symbol of a triac is shown in **fig. 9**. In the applications discussed here these devices act as solid-state ac relays. They have other uses, such as phase controlling and light dimming circuits, for example.

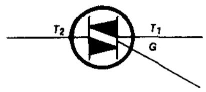

fig. 9. Symbol for a triac.

FCC test topics

The following Advanced FCC test topics are discussed in this article, but should be understood by Extra class license applicants also:

- diodes: zener, tunnel, varactor, hot-carrier, point-contact, PIN, light-emitting, neon
- transistors: junction, unijunction, power, FET
- *silicon controlled rectifier, triac*
- voltage regulator circuits, discrete and integrated
- voltage regulator with pass transistor and zener diode to produce a given output voltage.

For additional information, see *Electronic Communication* by Robert L. Shrader, McGraw-Hill Book Co., available from Ham Radio's Bookstore.

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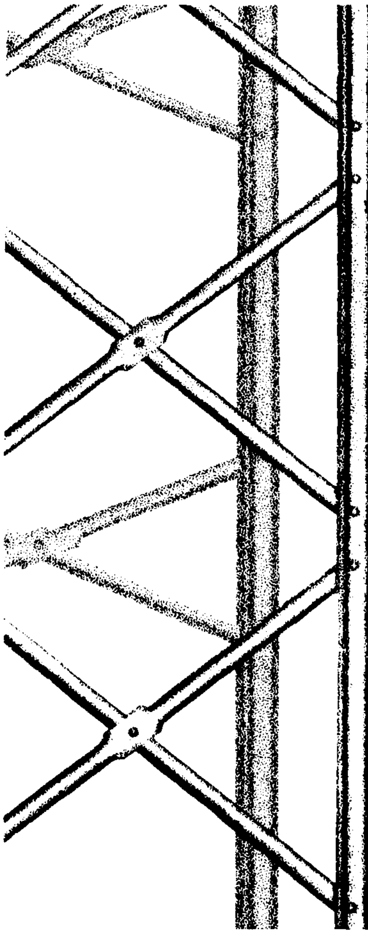

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systematic design of crystal ladder filters

Using theory and experimentation to remove some of the mystery of filter design

Reports have appeared on the use of crystal ladder filters for SSB.[1,2] However, there doesn't seem to be any general awareness that modern network synthesis provides a theory and tabulated data that can be used for an approximate systematic design of these filters. Such a systematic approach is described here together with some practical hints for applying it. Several examples illustrate the approach. A simple strategem is also introduced to permit the use of crystals that would otherwise be unsuitable for crystal ladder filters.

a bandpass case in modern filter theory

A great deal of practical information on modern filter theory has been published since WW II. For design of narrow bandpass filters the ITT handbook[3] gives tables that are very easy for an Amateur to apply, particularly in this day of the inexpensive handheld calculator. Both to simplify the discussion and to indicate how this material applies to crystal ladder filters, start with the configuration of **fig. 1**. This figure is viewed as a cascade of coupled resonators and is a generalization of the familiar double-tuned circuit. In the case taken here, coupling is by common capacitance. The design tables in reference 3 provide the coupling coefficients between adjacent resonators and the operating Q of the end meshes.

In the circuit of **fig. 1** there are n meshes, coupled by the $n\text{-}1$ capacitors $C_{1,2}$ . . . $C_{n\text{-}1,n}$. The first mesh

By Worthie Doyle, N7WD, 1120 Bethel Avenue, Port Orchard, Washington 98366

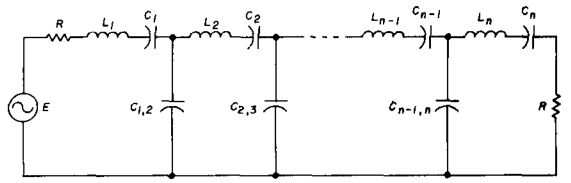

fig. 1. A bandpass ladder filter.

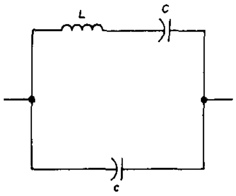

fig. 2. Crystal equivalent of fig. 1.

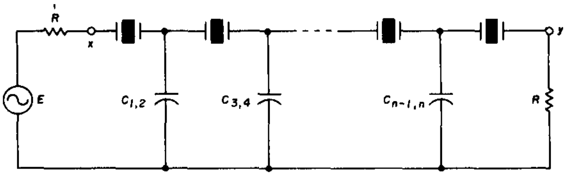

fig. 3. Simple crystal ladder filter.

consists of the source, its internal resistance R, L_1, C_1, and $C_{1,2}$. The second mesh consists of $C_{1,2}$, L_2, C_2, and $C_{2,3}$ and so on. The first and last meshes are loaded with the equal source and load resistances R. Each mesh is assumed to resonate at the center frequency, f_0, when its adjacent meshes are open circuited. This provides a practical tuning method, though it is not used in the crystal filters to be discussed. To tune a given mesh, put a large resistance in series with each of the series LC branches adjacent to that mesh. "Large" means large compared to R. This reduces the coupling to the given mesh, leaving its neighbors almost open. Set the source E to frequency f_0 and adjust the given mesh for maximum output. Do this for each mesh. Readers who have adjusted over-coupled, double-tuned transformers will recognize this as the series circuit analog of what is done with such transformers (you shunt one winding with a low resistance and adjust the other winding for maximum response at the center frequency).

The values of the network components are determined from the tabulated data[3] as follows. Let df be the 3-dB bandwidth of the filter. Let $Q_1 = Q_n$ be the Q of the end meshes, $Q = 2\pi f_0 L/R \approx 1/2\pi f_0 CR$. These Qs are related to normalized q, tabulated in reference 3, by

$$Q_i = q_i \frac{f_0}{df} \approx \frac{1}{2\pi f_0 C_i R} \quad (i = 1,n) \tag{1}$$

The coupling coefficients between two adjacent meshes are $K_{i,i+1}$. These coupling coefficients are related to normalized $k_{i,i+1}$, again given in reference 3, by

$$K_{i,i+1} = k_{i,i+1} \frac{df}{f_0} = \frac{\sqrt{C_i C_{i+1}}}{C_{i,i+1}} \tag{2}$$

$$(i = 1,2 \ldots n-1)$$

The tables in reference 3 give sets of k and q for several mathematical filter shapes — in particular the Chebychev for various values of passband ripple including zero (the Butterworth limiting case) and for up to seven meshes. These tables apply to a wider variety of filters and configurations than that of **fig. 1**, and the reference is well worth reading. Notice that the special case under consideration is symmetrical: $q_1 = q_n = q$ and $k_{1,2} = k_{n-1,n}$ and so on. Although not the subject of this article, the process above leads to useful LC sideband filters at frequencies up to about a hundred kHz.

To simplify further and to prepare for the crystal filter case, all the series branch Cs can be taken equal, say $C_1 = C_2 = \ldots = C$ and the meshes imagined tuned by small adjustments of L_i. Under these conditions the equations needed to carry out the design are, from **eqs. 1** and **2**:

$$R = \frac{df}{2\pi f_0 C q f_0} \tag{3}$$

and

$$C_{i,i+1} = \frac{f_0 C}{df k_{i,i+1}} \tag{4}$$

The tabulated q and the values of C, f_0 and df determine the source and load R. The tabulated $k_{i,i+1}$ for the chosen shape and the values of C, f_0 and df determine the coupling capacitors.

simple crystal ladder filters

In the neighborhood of its main natural frequency, a crystal looks like the circuit of **fig. 2**, where a small loss resistance can be ignored. The series and parallel resonant frequencies, f_0 and f_∞ are given by:

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \qquad f_\infty = \frac{\sqrt{\frac{1}{C} + \frac{1}{c}}}{2\pi\sqrt{L}} \tag{5}$$

It turns out that C is very much less than c, so the two frequencies are related by

$$f_\infty - f_0 \approx \frac{C}{2c} f_0 \tag{6}$$

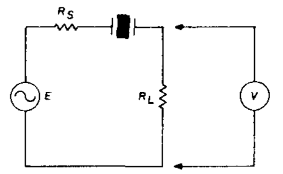

fig. 4. Measuring crystal C.

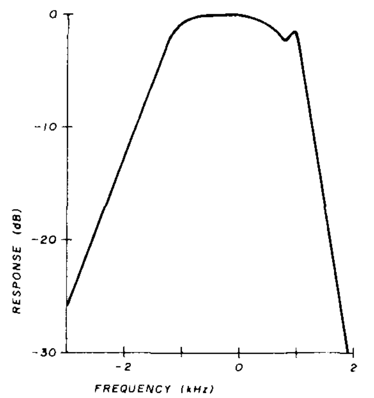

fig. 5. Response of four-crystal SSB filter.

For the time being, assume that f_∞ is sufficiently above f_0 to be well outside the passband of the filter to be designed. Under these conditions f_∞ and the crystal shunt c can be ignored. Then the configuration of **fig. 1** can be realized with crystals, as in **fig. 3**. (Design tactics when f_∞ is too close are discussed later.)

If all the crystals in **fig. 3** have the same f_0, the n meshes will still not be resonant at the same frequency. However, since the crystal C is very much less than $C_{i,i+1}$ in series with the crystal in each mesh, the mesh frequencies are not very different from f_0 and the meshes are almost correctly tuned, as required for the theory of **fig. 1**.

Since crystals come with L and C fixed, the crystal-filter design starts with C given, and **eq. 3** determines R, while **eq. 4** determines the value of the coupling capacitors. Manufacturers do not stamp the value of L, C and c on their crystals, but a good guess at C can be made from the setup shown in **fig. 4**.

If the crystal is driven by a grounded-base or a grounded-emitter amplifier, the source resistance, R_S, can be assumed to be the dc transistor load resistor. So that the crystal shunt c will not greatly affect the measurement, both R_S and R_L should be as small as convenient. Values up to one or two hundred ohms are adequate. The voltage across R_L is measured with an rf probe at the frequency of maximum response, f_0. Frequency is then varied above and below f_0 to find the frequencies where the voltage has decreased to $\sqrt{2}/2$ times the maximum first measured. The difference, df, between these two frequencies is then used in **eq. 7** to estimate C:

$$C = \frac{df}{f_0 2\pi f_0 (R_S + R_L)} \qquad (7)$$

This measurement on some crystals marked 5724 kHz gave $C = 0.026\ pF$. The shunt capacitance can be measured at a frequency well away from f_0 by any capacitance-measuring scheme. For these crystals it is about $c = 6.2\ pF$. These values in **eq. 6** imply that f_∞ is about 12 kHz above f_0, which is safe to ignore.

Only six of these crystals were on hand, so a four-crystal filter was built, with the other two set aside as carrier crystals. The resulting filter is unsuitable in a receiver for battling 20-meter behemoths but is quite usable in ordinary communications. It is perfectly adequate in a transmitter; and in any event, it illustrates how the theory is applied.

measurements

My setup for measuring C and also for measuring the constructed filters is a bit crude and can be easily duplicated, and probably improved, by other home brewers. It consists of a transistor Vackar oscillator with buffer amplifiers to drive the filters and a diode probe connected to a very high-resistance dc voltmeter to measure output. Frequency is read on a home-made counter. Resulting measurements from the first design are used to re-estimate C, leading to a closer and usually final design.

experimental SSB filter

For the four-crystal filter with $f = 5724\ kHz$, choose $df = 2.5\ kHz$ and the 1 – dB ripple Chebychev shape. The table[3] gives $k_{1,2} = 0.638 = k_{3,4}$; $k_{2,3} = 0.546$ and $q = 2.21$. **Eq. 3** gives $R = 467\ ohms$, and **eq. 4** gives $C_{1,2} = C_{3,4} = 93\ pF$, and $C_{2,3} = 109\ pF$. This filter was assembled with $R \approx 470\ ohms$ and capacitors measured to about 5 percent from the junk box. My measurements indicated that the filter had a 3-dB bandwidth of about 1650 Hz and a maximum ripple of about 1.4 dB near the top of the passband. Instead of the four peaks of the theory, there was one peak near the top end and a very broad single peak over the rest of the passband.

If you believe in the theory, the experiment above tells you that the estimate for C from **eq. 7** is too high. I therefore guessed C to be about 0.005 pF less, namely 0.019 pF and recalculated $R = 639\ ohms$, $C_{1,2} = C_{3,4} = 68\ pF$, and $C_{2,3} = 80\ pF$. With the second round of coupling capacitors selected from the junk box and the source and load R set near 640 ohms, the measured bandwidths were 2400 Hz at 3 dB, 2650 Hz at 6 dB and about 4740 Hz at 30-dB down. My rf measurements are certainly not reliable beyond 30-dB down. This filter has a 30- to 3-dB shape factor of about 2.0. For the 1-dB Chebychev characteristic it should be about 1.75, as obtained from the curves in the ITT handbook. However, this filter seems quite satisfactory, and its response is plotted in **fig. 5**. Since the coupling capacitors can be measured only to 2 or 3 pF, further work is not indicated.

Incidentally, as a general observation, changing the source and load resistors from the calculated values has only a small effect on the bandwidth but does change the passband ripple and skirt slopes. For lower R the ripple increases and the skirts steepen; for higher R the reverse results. Small changes in R can be used to make some final improvements if desired.

The filter characteristic plotted in **fig. 5** has one dip of 1.4 dB near the top of the passband. The loss of this filter, measured as the ratio of voltages at **x** and **y** in **fig. 3**, is about 2.6 dB (1 volt out for 1.35 volts in).* Considering the known deviations from the theoretical model, the agreement seems reasonably satisfactory. The meshes are not tuned to quite the same frequency, both because of the variations from crystal-to-crystal and the effects of the different coupling capacitances in each mesh.

analysis

The shunt c will cause the filter, even inside the passband, to deviate from that which would be obtained in the absence of c. The reactance of a crystal at any frequency below f_∞ is always algebraically greater than the reactance of the series LC branch alone, except where they touch at f_0. Between f_0 and f_∞ the crystal has a greater inductive reactance and rate of increase than the LC branch alone. Below f_0 the crystal has capacitive reactance, whose magnitude is less than that of the LC branch alone. The rate of increase of reactance below f_0 is also less for the crystal than for just its LC branch. These differences, though small near f_0, can be expected to cause deviations in filter shape from the theoretical curves for the filters of **fig. 1**. Furthermore, whatever these deviations are, they should have opposite characteristics above and below f_0. From the example of **fig. 5** it appears that the effect is to increase ripple above f_0 and to decrease it below f_0. At f_0 the shunt c has no effect.

If the crystals were shunted by an inductance that cancels c at f_0, I would expect passband shapes to agree with the theory for **fig. 1**. This is because the high reactance of the parallel-resonant circuit easily persists over the passband, making the net crystal reactance that of the series LC in this region. I hope to try the experiment some time when suitable coils can be found or wound. However, I would not use such a filter in practice, because values of f_∞ closer to the passband give better filter shape factors — a subject discussed in a later section.

*The editor and author are aware of the concept of insertion loss. The author finds the ratio of voltages at points x and y easier to measure and just as useful.

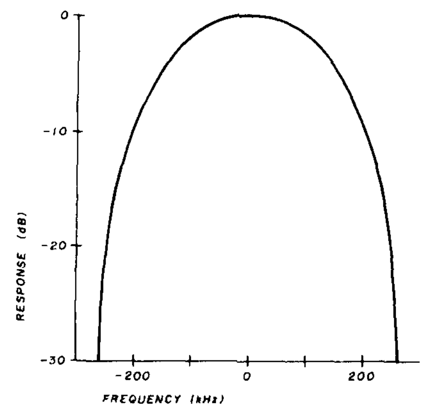

fig. 6. Response of four-crystal CW filter.

experimental CW filter

As a second example, here is a narrow CW filter, also using four crystals. These are type CR1A/AR marked 5910 kHz. Series C measures 0.0023 pF and shunt c about 4.5 pF. For this example choose $df = 200\ Hz$ and the Butterworth shape. The appropriate table in reference 3 gives $q = 0.766$ and $k_{1,2} = k_{3,4} = 0.840$; $k_{2,3} = 0.542$. **Eq. 3** determines $R = 517\ ohms$, and **eq. 4** determines $C_{1,2} = C_{3,4} = 80\ pF$ and $C_{2,3} = 124\ pF$. When this filter was built, the bandwidth turned out to be about twice what was expected. The series C was therefore re-estimated to be 0.0046 pF. The recomputed coupling capacitors are $C_{1,2} = C_{3,4} = 162\ pF$ and $C_{2,3} = 251\ pF$. The recalculated load resistances are 256 ohms. This filter was built with 160 and 250 pF from the junk box assortment of mica capacitors. The characteristic for this filter appears in **fig. 6**. The 3-dB bandwidth is a little under 250 Hz and a third trial seems unnecessary for my application. The 30-to 3-dB shape factor is about 2.0. The theoretical value for a Butterworth filter is about 2.4, so this filter has slightly steeper skirts than expected. As noted above, the skirts could easily be spread slightly and the characteristic made closer to the Butterworth shape by raising the terminating resistances to perhaps 400 ohms.

comments on CW filters

For a given bandwidth and number of resonators, the Chebychev design contains a trade-off between passband ripple and skirt steepness. For speech filters high ripple and steep skirts are probably preferable. As ripple is decreased so are overshoot and ringing, so it seems likely that for a narrow CW filter the Butterworth shape would be more satisfactory.

As a point of information, the ITT handbook also contains design data for Bessel filters. Bessel filters produce minimum distortion of pulse envelopes for a given bandwidth but have even broader skirts than Butterworth filters. The handbook contains theoretical curves from which shape factors can be read. The 30- to 3-dB shape factors for 1-dB Chebychev, Butterworth and Bessel filters using four resonators are 1.75, 2.4, and 3.5 respectively.

Anyone who has read advertisements for CW filters must have noticed that they have significantly higher shape factors than the same manufacturer's SSB filters. The above discussion may help to explain this observation. It seems to be customary to quote 60- to 6-dB shape factors for filters. Since I cannot measure reliably below about 30 dB, shape factors for 30 to 3 dB will be given when comparing results to theory. Readers who must have shape factors for other levels can easily estimate the theoretical values from the curves in the ITT handbook.

effect of crystal shunt capacitance

Hardcastle[2] points out that some crystals are unsuitable for the simple SSB filter of **fig. 3**. These are crystals for which f_∞ is too close to f_0. The passband spreads slightly above and below f_0. If the crystal f_∞s are too close, the filter's high-frequency cutoff will come "early" and the available bandwidth may be too narrow for SSB. I ran into this problem trying to use some of the crystals in the previously described CW filter for a six-crystal SSB filter at 5910 kHz. The recipe of the previous section was used but no amount of struggling with coupling capacitors or terminating resistances availed to produce a passband wider than about 1000 Hz. Taking the best estimates of crystal parameters, $C = 0.0046\ pF$ and $c = 4.5\ pF$, we find from **eq. 6** that f_∞ is about 3 kHz above f_0. Apparently this is too close. Incidentally, my CR1A/AR crystals are plated squares, supported on wires at diagonally opposite corners and are not the compression-mounted CR1AR crystals Hardcastle notes to be unsuitable. Nevertheless, my 5910-kHz crystals are equally unsuitable!

inductance coupling

An interesting experiment was made to test this notion about my crystals. First, notice that the mesh frequencies will be higher than the crystal f_0, since the crystal is in series with the coupling capacitors. This aggravates the effect of a too-low f_∞. Now, coupled resonator filters can be coupled just as well by common inductance as by common capacitance. The series inductances will lower the mesh frequencies, moving the passband away from the crystal f_∞. I happened to have several small encapsulated 26-μH chokes whose reactance at 5910 kHz is close to that of the appropriate coupling capacitors. Without changing anything else, the coupling capacitors in an unsuccessful trial filter that was 1-kHz wide were replaced by these chokes, whereupon the filter bandwidth went to about 4500 Hz. This verifies the cause of the trouble and suggests the two tactics to be discussed in the following sections. Common inductance coupling can apparently move f_0 sufficiently to make these crystals usable, and I plan some more careful trials. The value of coupling inductance could be directly calculated from an equation similar to **eq. 4**. An alternative is to calculate coupling capacitances from **eq. 4** and then replace them with the inductances with which they resonate at f_0. A minor problem arises because crystal shunt c and the coupling inductances form a highpass filter some MHz above the passband. A few high-Q LC circuits at f_0 should eliminate that response.

use of series inductance

Since series inductance lowers the f_0 of a crystal while leaving f_∞ fixed, why not just replace the simple crystal ladder filter by the configuration of **fig. 7**? The junk box happened to contain a number of small encapsulated rf chokes that had been measured as 62 μH shunted by 1.8 pF, so one was put in series with each crystal and the filter response roughly checked by watching the output voltage at point **y** in **fig. 7** as the VFO was tuned through the passband. Coupling capacitors were some left from one of the earlier trials with these six unsuitable crystals. For load resistances of 200, 300, 400, and 1000 ohms the 3-dB bandwidths were 2110, 1970, 1740, and 1670 Hz respectively. Except for the case of 1000-ohm terminations, all the shapes were reasonably flat on top. Filter loss varied from 13 dB with 400-ohm terminations to 19 dB with 200 ohms. This loss is computed from the ratio of voltages at points **x** and **y** in **fig. 7**.

For my projected use of this filter, a loss of 19 dB seemed excessive. Filter output voltage would be only a tad over one-tenth of the input voltage. For this reason no response curves were plotted. However, the shapes were good, and if a loss of this magnitude can be tolerated, the configuration is worth considering as a way to use crystals that are unsuited for use in the simpler configuration of **fig. 3**. To build such a filter I would determine the coupling capacitor

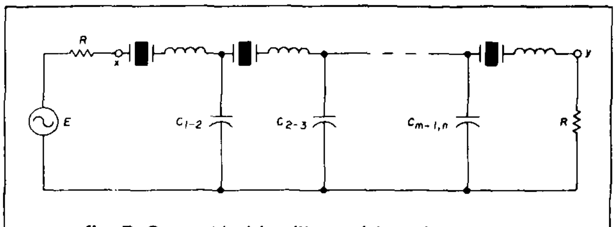

fig. 7. Crystal ladder filter with series coils.

values as for the simple **fig. 3** configuration but would use a resistance slightly lower than the calculated terminating resistance. The latter variation should compensate somewhat for the loss in the chokes.

use of shunt inductance

To introduce this point, recall both the cause of f_∞ and its effect on the filter. The cause, of course, is c of **fig. 2**. The effect is not only to interfere with the filter's upper frequency limit but also to produce an n-fold null in the filter's response, or a series of closely spaced nulls just above the filter's passband. This is the result of parallel resonances in the signal path. These nulls have the effect of steepening the high-frequency cutoff, as is clearly shown in Hardcastle's curves[2] and in **fig. 5**. This makes the simple crystal ladder filter better for LSB than USB.

Considering its source, f_∞ can clearly be moved by shunting the crystal with reactance. More capacitance would only move f_∞ closer to f_0. The effect of shunt inductance depends on the magnitude of the inductance. If the inductance is greater than the value required to resonate c at f_0, then the effect is to move f_∞ up, thus converting unsuitable crystals to suitable ones if moved far enough. To see why this should be so, notice that such an inductance cancels part of the shunt c at f_0. This inductance, with the cancelled part of c, is parallel resonant at f_0 and does not affect the operation of the series LC branch of the crystal near the passband. The residual, uncancelled part of c is, of course, smaller than the original c and combines with the series LC to produce a higher f_∞ than the original.

If the shunt inductance is less than the value required to resonate at f_0 with the shunt c, then the effect is to move f_∞ below f_0. One way to see this is to imagine the shunt inductance broken into two greater inductances in parallel, one of which is the value required to resonate at f_0 with c. This latter combination, as in the previous case, does not disturb the crystal operation near the passband, while the remaining portion of the shunt inductance combines with the series LC to produce an f_∞ below f_0. If f_∞ is pushed sufficiently below f_0, an unsuitable crystal can again be made suitable.

Useful values of shunt inductance will be on the order of the inductance required to resonate c at f_0. Let x be the inductance that resonates at f_0 with c. Let $x + dx$ be the value of shunt inductance actually used. Note that dx is not required to be small compared to x and may be positive or negative. The new f_∞ is then determined approximately by

$$f_\infty - f_0 = \frac{C(x+dx)f_0}{2c\,dx} \tag{8}$$

or, solved for the unknown dx:

$$dx = \frac{x}{\frac{f_\infty - f_0}{f_0} \bullet \frac{2c}{C} - 1} \tag{9}$$

Real coils have parasitic shunt capacitance that causes a slight difficulty in **eq. 9**. The values of inductance that will be called for with high-frequency crystals, roughly in the range 20 to 200 μH, can be expected to have shunt capacitances from 1 to 5 pF. An estimate of this shunt capacitance must be added to the crystal shunt capacitance to obtain the c and x used in **eq. 9**, resulting in a certain amount of imprecision and trial. Nevertheless, good estimates of the required shunt inductance are possible.

an example using eq. 9

Suppose for my unsuitable 5910-kHz crystals a new f_∞ is desired 10 kHz below f_0. Assume the coil will have 3.5 pF shunt capacitance. Using the crystal values of 0.0046 and 4.5 pF, we have $C = 0.0046$ and $c = 8$ in **eq. 9**. The inductance resonating with 8 pF at 5910 kHz is $x = 91\ \mu H$. Inserting all this into **eq. 9**, we get $dx = -13$. The inductance to use is therefore $x + dx = 78\ \mu H$. If, on the other hand, we want f_∞ 12 kHz above f_0, the result is $dx = 15$, and 106 μH is the value to use in the circuit. In the actual case 3.5 pF may overestimate the shunt capacitance of a 78-μH coil, so that slightly greater inductance might finally be used.

Eq. 9 will usually be used in two ways, neither of which suffers much from imprecise estimates of coil shunt capacitance. The first use of **eq. 9** is in moving crystal f_∞ sufficiently above or below f_0 to make unsuitable crystals usable and to produce the sharper cutoff on the desired side of the passband. The other use is to produce nulls at particular frequencies, as discussed in a later section. In that case, coils will be used that are adjustable in the neighborhood of the value calculated from **eq. 9**, and the nulls will be set to the desired frequencies whatever the actual shunt c.

To keep the record straight, when a crystal is shunted with inductance the combination has two values of f_∞, one below and one above f_0. The approximation (**eq. 8**) applies to the one close to f_0.

Having no small handy chokes near the above sizes, I took the 62-μH chokes of the previous section's experiment and shunted them across the six crystals, producing the filter of **fig. 8**. From the discussion above it is clear that with all f_∞ below the passband, the filter should have its steeper cutoff at the low end and thus be better suited for USB, which is my application. The curve for the filter of **fig. 8** is given in **fig. 9** and indeed has a very sharp low-frequency cutoff. The shape is a good way from the 1-dB Chebychev, for example, yet it is quite useful. The bandwidths at 3, 6, and 30 dB are 2300, 2700, and 3900 Hz respectively; the 30- to 3-dB shape factor is about 1.7.

The subpeak just above low-frequency cutoff is 1.2 dB down from maximum response, and the trough just above it is 1.4 dB down from maximum response. If the carrier were placed at the 30-dB-down point, the small subpeak would correspond to an audio frequency of about 250 Hz. I'm sure this filter can be improved further, but lethargy triumphed.

It is interesting that when f_∞ is moved just outside the low end of the passband, the effect of shunt inductance on the passband shape is just the opposite of that of crystal shunt c on the simple filter of **fig. 3**.

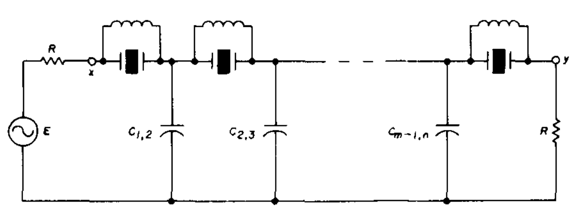

fig. 8. Crystal ladder filter with shunt coils.

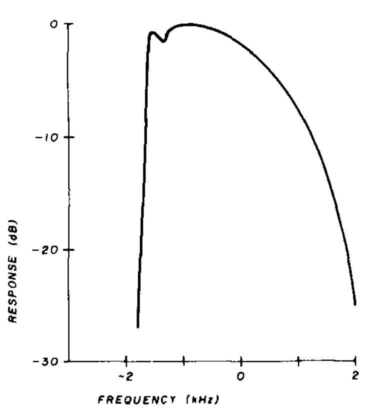

fig. 9. Response of six-crystal shunt coil SSB filter *using "unsuitable" crystals.*

Ripple increases at the low end and decreases at the high end. Without going into details, the deviations in the reactance of the crystal branch with shunt inductance from the reactance of the series LC branch alone are opposite from the deviations with crystal shunt c.

Terminating resistors used with this filter were 1000 ohms. The loss, as measured in the previous section, is 2 dB at the top of the broad peak. This shape and loss are so satisfactory for my use that further tinkering is not contemplated despite the departure from book shape. I calculate the new f_∞ to be about 2.5 kHz below f_0, taking into account the choke shunt capacitance of 1.8 pF. The six nulls crowd toward the low end of the passband just as the original six crowded toward the high end and prevented achievement of a bandwidth over 1 kHz. The coupling capacitors used in this filter are smaller than the theoretical values for a 2.5-kHz bandwidth, and the passband is shrunk at the low end.

Coils shunting the crystals introduce a minor hazard of which to be aware. Some MHz below the passband these coils and the coupling capacitors form a lowpass filter. Response there must be reduced by other means. Ordinarily there will also be some LC circuits at the filter frequency, and these should solve the problem. Also, to avoid unnecessary loss, the filter should be matched to the devices between which it operates. The hazard can be avoided by making these matching circuits high-Q tapped tanks or high-Q L networks (Q of an L network can be increased by replacing the series element by a series LC circuit or the shunt element by a parallel LC circuit.) Of course, this hazard will not occur if your crystals can serve in the simple circuit of **fig. 3**.

adjustable zeros

The discussion above of f_∞ and its manipulation suggest the possibility of using adjustable coils across each crystal and setting half the f_∞ to specific frequencies above the passband and the other half to specific frequencies below the passband. This can make a great improvement in shape factor and produce performance close to that of elliptic function filters. If the passband ripple is fixed and the minimum stop-band attenuation is fixed, then elliptic function filters give the fastest possible falloff from the band edge to the specified minimum stop-band attenuation for a given number of resonators. Unfortunately, the ITT handbook does not give data to determine element values, partly, I suspect, because too many cases would need to be listed. However, curves are given for the achievable performance of various combinations. Using these as a guide to what is possible, a dedicated experimenter should be able to achieve useful results by trial, at least in a modest case like

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More Details? CHECK—OFF Page 100

four crystals and four adjustable shunt coils. I would start by using the values of coupling capacitors for something like the 1-dB Chebychev case and the corresponding terminating resistance. For this kind of tinkering, one should have a swept oscillator and scope display, preferably logarithmic — equipment unavailable to most of us. Amateurs who work for tax-supported or nonprofit organizations may be able to do some of this tinkering after hours at their places of employment.

other observations

A side issue that needs to be mentioned is the use of extra shunt capacitance across the source or load. These have appeared in some of the references, perhaps by analogy with some image-parameter filter designs. They are essentially unnecessary. They have two effects. First, as with the coupling capacitors in the network, they slightly and insignificantly increase the frequencies of the source and load meshes. Second, they cause the shunt load resistance used on the filter to look like a different resistance in series with the crystal. Recall that the equivalent series resistance is what determined the Q of the end meshes in the design. This parallel to series RC transformation could be part of a matching arrangement, but I would forget it.

summary

Although there has still been some element of trial in the filters constructed, it seems to be mainly attributable to uncertainty in the measurement of the crystal series capacitance. Few trials were needed, generally two, and the terminating resistances and coupling capacitances are theoretically related in a simple way to this one parameter.

The theory reviewed provides a simple, systematic approach to the design of crystal ladder filters. Finally, the use of shunt inductance has been shown to solve the problem of using "unsuitable" crystals and to allow for predetermined nulls in the stop band. These nulls can be used to improve the rolloff at both band edges.

I hope this note will remove some of the mystery from the design of crystal ladder filters and encourage other home brewers to design and build their own.

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1. Peter J. Hampton, G4ADJ, "Using TTL ICs in the SSB Equipment," *ham radio*, November, 1975.
2. J.A. Hardcastle, G3JIR, "Some Experiments with High-Frequency Ladder Crystal Filters," *QST*, December, 1978.
3. *Reference Data for Radio Engineers*, 4th edition, IT&T Corporation, 1956. More recent editions will also contain the curves and tables required.

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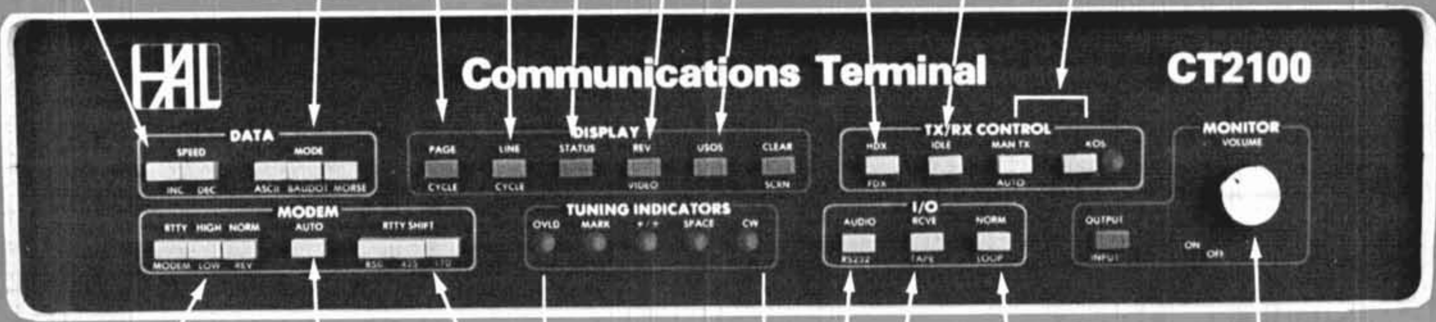

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- Wide receiver dynamic range. Junction FETs in the balanced mixer, MOSFET RF amplifier at low level, and dual resonator for each band.
- Variable bandwidth tuning (VBT). Varies IF filter passband width.
- Notch filter (high-Q active circuit in 455-kHz second IF.
- IF shift (passband tuning).
- Built-in digital display (six digits, fluorescent tubes), analog dial, and display hold (DH) switch.
- Noise-blanker threshold level control.
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Auxiliary power system designed for mobile use for the ICOM IC2AT 2-meter handheld transceiver. At right is the homebrew regulator, built around the LM317 IC. Input to the regulator (top) is an accessory plug that connects to a socket beneath the car dashboard (or you could connect it to the car cigar-lighter receptacle). Miniplug on the output end connects to a BP4 battery case, which contains six rechargeable nicads. The small addition on the bottom of the BP4 battery case is described in the text. Note the hole in the leather carrying case, which was made for the external power plug.

two-way power for the IC2AT 2-meter handheld

Running out of battery power on the freeway? Here's a neat solution using a homebrew regulator

I was in rush-hour traffic on the interstate, with one hand on the steering wheel and the other holding my new ICOM IC2AT transceiver, when suddenly the transmit-indicator light went out, indicating that my nicads had run down. What to do? Try to remove the radio from its leather case and change battery packs while driving? Impossible! Pull off the road and change batteries? Too much traffic! Only one choice left — I made a quick apology to the people with whom I was in contact and shut down.

This situation occurred two days after I had purchased my IC2AT, and it made me realize that 40 minutes of rag chewing twice a day during my drive to and from the office was just too much for my nicads to take. I immediately began thinking of a better way to enjoy the versatility of this neat little 2-meter rig.

about the IC2AT

The IC2AT hand held has become an instant success with Amateurs in my area. With its compact size, high-quality construction, excellent operating reports, and low price it's hard to beat. I bought one with the objective of having a pocket-size 2-meter rig that I could use in my car while commuting to work; from my desk in my office as well as on my boat; at hamfests; and on occasional business trips to other areas of the country.

By Gil Weiss, WB3JJF, 3004 Hallowell Court, Bensalem, Pennsylvania 19020

Rear view of the modified BP4 battery pack showing charger jack.

A spare battery pack was required, and I decided on the BP4 battery case, which I loaded with six AA-size rechargeable nicads. This spare power pack lasts much longer than the standard 250-milliampere-hour battery that comes with the rig. The power output of the radio is slightly lower, however, because of lower voltage from the six 1.25-volt nicads. But even armed with two batteries, I still had mobile power problems, because I couldn't remove the unit from its protective leather case and change power sources while driving.

operating problems

Purchasing the dc regulator offered by ICOM appeared to be the solution to my problem. The regulator is built into a slide-on battery case and decreases the 13.8-volt auto voltage to the required 8.4 volts specified in the ICOM instruction manual. But — and there always seems to be a *but* or *however*, as nobody seems to make the perfect rig or accessory — the jack that's used to plug the regulator base into the auto electrical system is located on the back of the unit, directly behind one of the big metal snaps on the rear of the leather case.

Even if you drill out the snap (which would not do your case much good) or used the rig out of the case, laying it on your car seat or hanging it on your dashboard, you still have the problem of the plug and wire coming out of the rear of the radio. When I got to work I'd have to remove the unit from the leather case, take off the regulator, put on a battery pack, replace the unit in the case — and then repeat the entire procedure when leaving work for home. A better solution was needed.

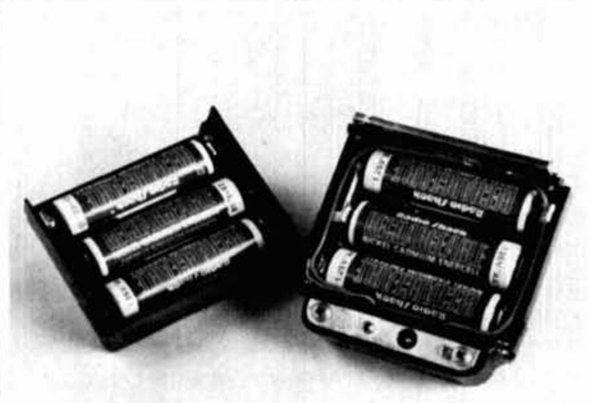
the two-way supply

The product of my imagination is shown in the photos. I built a voltage regulator circuit using an LM317T variable-regulator chip in a case measuring 3-1/4 x 2-1/8 x 1-1/8 inches (8.3 x 5.4 x 2.9 cm). One end plugs into the auto electrical system either through the cigar lighter or, as in my case, through an accessory plug wired under my dashboard. The wire from the output end terminates in an 1/8-inch (3-mm) miniplug, which goes to the main part of my invention — a BP4 battery case which holds six AA nicads and which has a small addition on its base that accepts the external power plug from the regulator.

When you plug in the regulator the normally closed miniplug jack disconnects the internal batteries, and when you remove the miniplug, the batteries are once again connected and ready to provide portable power. The leather case is included in the photo to show where I made the hole for the external power plug on the lower front of the radio.

For added convenience, a charger jack is also built into the battery case addition, which allows using the BCU-25 wall charger that comes with the radio or any other 10-15-volt 50-mA dc converter/charger. You don't need to use the ICOM drop-in quick charger for the BP4 pack, as stated in the IC2AT manual.

Another photo shows the rear of the modified BP4 pack, which includes the charger jack. The addition to the battery pack case was cut from a small box, just like that used for the regulator circuit. I cut and epoxied it to one side of the BP4 pack so the case could be opened and closed to install or remove batteries. This piece of the box contains a normally closed miniplug jack, which does the switching from internal to external power, plus a coaxial charging jack.



Interior of the battery case with nicads installed.

Those who complete this easy project will be rewarded with a great little 2-meter rig which, after operating indefinitely from your auto electrical system, can be quickly removed by disconnecting your mobile antenna and unplugging your external voltage regulator. Just put your rubber duck back on and you're instantly ready to operate portable. All this can be done while keeping your radio safe in its leather case. Additionally, with your dc regulator you can run the rig from your home on any 12-volt power supply, or an old auto battery, which could provide communications capability during power outages.

construction

The regulator schematic is shown in **fig. 1**. I built the circuit on a piece of perf board cut to fit inside the box. The regulator chip and heat sink were placed on the bottom side of the board; all other components were placed on top. (All components are available at Radio Shack stores except for the 12-volt, 5-watt zener, which was included for over-voltage protection of the rig.) If a spike greater than 12 volts should occur, the zener will shunt the circuit to ground and blow the fuse.

A subminiature slide switch and a green LED were also included, since the regulator should always be turned off when inserting the output plug into the rig. (These plugs easily short out for an instant while being plugged in or out of their jack.) Wire the tip of the miniplug for 8.4 volts, and upon completion of the regulator circuit, adjust the 5k variable resistor for 8.4 volts output before installing the top of the box.

Fig. 2 shows the parts layout and wiring diagram for the battery case addition. Unsolder the orange (battery positive) wire inside the BP4 case from the positive connection at the top of the case and reroute it into the addition. Connect it to the enclosed jack.

Wire routing is done through two holes drilled in the two square indentations on the bottom of the BP4 case. A new wire goes from the positive side of the enclosed miniplug jack to the slide on the connector at the top of the battery case, where the orange lead was disconnected. Route another wire connecting the negative terminal on the miniplug jack and the negative side of the charging jack to the negative connection inside the BP4 case. Wire the charging jack in parallel with the battery circuit as shown. This completes all electrical work.

The mechanical part starts with cutting off a portion of the second box to the dimensions shown in **fig. 3**. The inside space must be just deep enough to fully enclose the miniplug jack. You can't make this addition too deep, or the whole unit won't fit into the leather case. I used a fine hacksaw and carefully sanded the piece of box to final dimensions with No. 400 silicon paper.

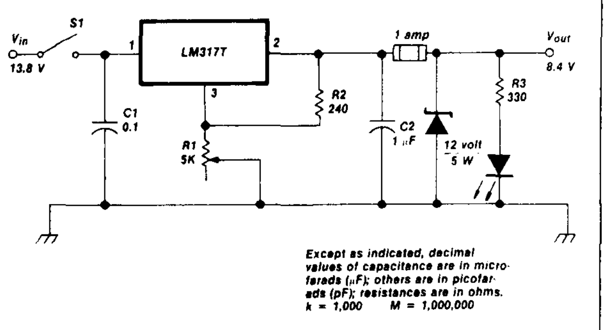


fig. 1. Schematic diagram of the regulator circuit. Zener provides protection against voltage spikes.

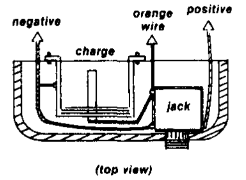


fig. 2. Parts layout and wiring for the battery-case addition.

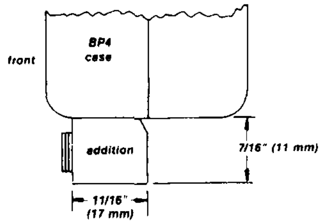


fig. 3. Surgery necessary to include the additional box to the BP4 case. See text for correct method of adding this unit.

Some interior trimming with a knife was necessary to make room for the wires. The charging jack also required some filing to fit properly. I also lightly sanded the entire bottom of the case to ensure a good bond with the epoxy, which will hold things permanently in place. Either the quick-setting (5-minute) or regular resin/hardener epoxy may be used for this job.

The metal side of the addition was carefully drilled to fit the charging jack, and the edge was beveled with the silicon paper on a block so that the case could easily be opened and closed.

A little patience and care with your saw, file, and abrasive paper will produce a result that will greatly add to the operating convenience of the IC2AT.

ham radio

ham radio TECHNIQUES

Bill Orr W6SAI

One of the interesting challenges of writing a monthly column is that you "write behind" the publication date by about three months. That is, this piece will appear in the February, 1982, issue of *ham radio*, but it is being written the first weeks of October, 1981. As of this writing, the future of the new "30 meter" Amateur band (10.1 to 10.15 MHz) is somewhat uncertain. Assigned at the World Administrative Conference of 1979 to the Amateur service on a secondary basis, this band could possibly be open to Amateurs worldwide by January 1, 1982. Indeed some countries have already announced the band as open to hams as of that date.

The situation is a little more complex in the United States, as the Senate has not yet ratified the treaty containing the provisions for the new band. And no one knows what the FCC has in mind for the band so far as Amateur occupancy goes. The latest scuttlebutt I hear is that a decision will be made in "early spring, 1982." That assumes, of course, that no hang-ups pop up along the way.

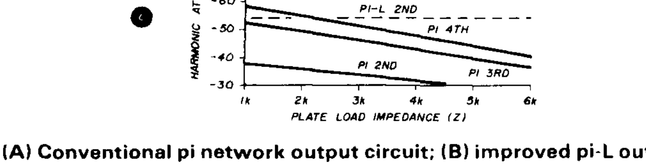


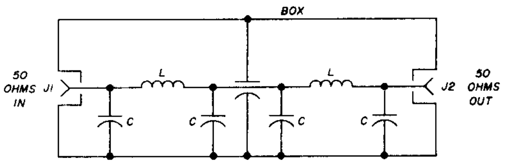
fig. 1. (A) Conventional pi network output circuit; (B) improved pi-L output circuit; (C) attenuation curves for pi and pi-L networks as a function of plate load impedance.

In any event, the band is in the offing and many Amateurs are wondering if their equipment will work on the new band and what problems, if any, they may encounter in firing up on the new frequencies.

Some of the more modern exciters have a bandswitch position for 30 meters and many older pieces of equipment (such as the Collins S-line and the Drake T4-R4 combos) provide tuning information and crystal data for the 10-MHz band. Thus many hams are all ready to go now, and others with older gear can be on the band by buying a new conversion crystal or two and retuning their equipment as outlined in the instruction manual.

watch out for transmitter harmonics

Remember that the harmonics of a 10-MHz signal do not fall in any



band	C	L
160	1700 pF	4.2 μH. 22 turns No. 16 (1.3 mm) AWG solid, 1" dia. × 2" long (2.5 × 5 cm)
80	850 pF	2.1 μH. 13 turns No. 16 (1.3 mm) AWG solid, 1" dia. × 1-3/4" long (2.5 × 4.4 cm)
40	440 pF	1.1 μH. 8 turns No. 16 (1.3 mm) AWG solid, 1" dia. × 1" long (2.5 × 2.5 cm)
30	330 pF	0.8 μH. 9 turns No. 16 (1.3 mm) AWG solid, 3/4" dia. × 1" long (1.9 × 2.5 cm)
20	220 pF	0.55 μH. 7 turns No. 16 (1.3 mm) AWG solid, 3/4" × 7/8" long (1.9 × 2.2 cm)
15	160 pF	0.41 μH. 8 turns No. 16 (1.3 mm) AWG solid, 1/2" dia. × 1" long (1.3 × 2.5 cm)
10	110 pF	0.3 μH. 6 turns No. 16 (1.3 mm) AWG solid, 1/4" dia. x 3/4" long (0.6 x 1.9 cm)

fig. 2. Design data for half-wave filter.

Amateur band. The burden of harmonic suppression to protect other services falls upon the Amateur operator.

Unfortunately, most commercial Amateur transmitters employ a simple pi-network output circuit, or a broad-band coupling transformer (in the case of a solid-state rig). The harmonic rejection of these circuits is marginal at best and can be insufficient to suppress bothersome harmonics that might interfere with services in the 20-MHz region. Particularly in the case of the tube-style transmitters with pi-network output circuit, although they can be retuned quickly to the 10-MHz band their pi-network design is optimized for operation in the nearby Amateur band. Retuning the network to the new band, even though the instruction manual tells the operator how to do it, may mean that the suppression ability of the network is seriously impaired.

As an example, suppose you have a transmitter that you are retuning from 7 MHz to 10 MHz (a common case). According to the manual, the bandswitch is left on the 7-MHz (40-meter) position, but the controls are tuned to new reference points for 10 MHz as given in the manual. You are now on 10 MHz! But what happened to the harmonic suppression of the output circuit? The manual doesn't say.

Harmonic suppression ability of pi and pi-L networks is illustrated in **fig. 1**. Let's compare the harmonic suppression of the two configurations. Take the case of a pair of 6146s operating at a plate potential of 800 volts and a peak plate current of 220 mA. The approximate plate load impedance in this case is about 2140 ohms. Second harmonic rejection with the conventional pi-network is about −35 dB, and third harmonic rejection is about −47 dB. This marginally meets FCC requirements for good engineering practice.

Now note that under the same operating conditions, second harmonic rejection by the pi-L network is about −52 dB and third-harmonic rejection is about −65 dB. A great improvement!

But how many Amateur exciters or transceivers make use of a pi-L output network? I'm afraid that most Amateur equipment designs are straight out of the Stone Age in this respect. A pity, since factory conversion to a pi-L design is inexpensive and nearly 20 dB of extra harmonic rejection can be had for pennies.

And the harmonic problem can grow worse when the exciter is hooked to a linear amplifier. Have you ever heard the harmonic of a SSB signal? It is the darndest sounding, Donald-Duck-type of gabble. But enough intelligence exists in the speech for a sharp-eared FCC monitoring station to establish your call letters.

simple add-on harmonic-suppression circuit

There's no need to revamp your station equipment to reduce transmitter harmonics to a reasonable level. You can add a high-frequency harmonic filter between the equipment and the antenna. The half-wave filter design shown in **fig. 2** will do the job.

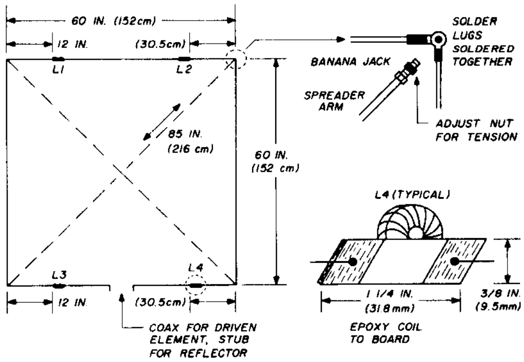


fig. 3. Mini-quad assembly. Mini-quad loop is 60 inches (152 cm) on a side for each element. Coils are placed 12 inches (30.5 cm) from the ends of the horizontal wires. For center frequency of 28.6, L1 or L2 are 20 turns, L3 and L4 are 22 turns. Use No. 22 enamel wire on T-50-6 Amidon form (yellow). Iron-powder core is 0.5 inch (12.7 mm) outside diameter, 0.303 inch (7.7 mm) inside diameter, permeability 8. For 29.6 MHz, L1 and L2 have 19 turns, L3 and L4 have 21 turns. Driven element to reflector spacing is 50 inches (127 cm). See photos for assembly details.

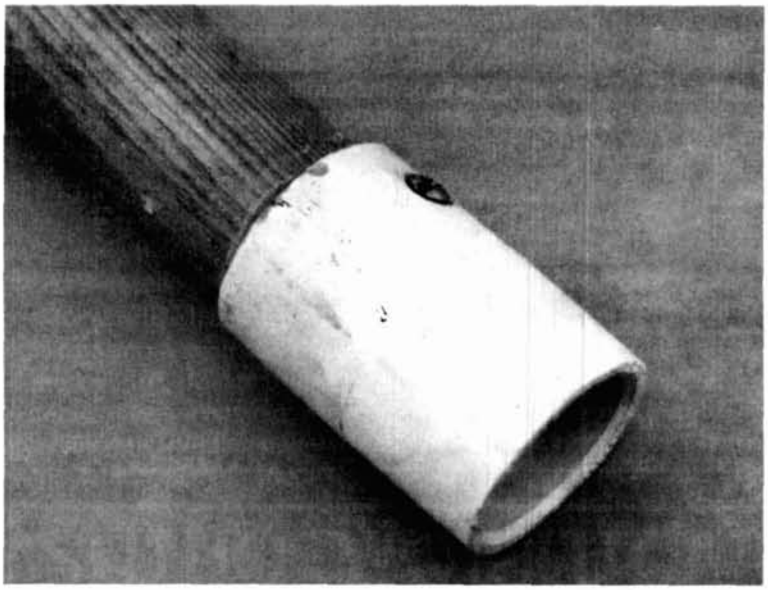


fig. 4. Wood boom has PVC plastic pipe union driven on it and screwed in position. The plug of the cross-arm support slips into open end of the union. Boom is coated with varnish after completion to prevent the wood from warping.

This unit has a cut-off frequency somewhat above the band in use; it provides about 30 dB attenuation to the second harmonic of the transmitter and approximately 48 dB attenuation to the third harmonic. Component values are given for all high frequency Amateur bands, but the one of greatest interest at the moment is the model for the 10-MHz band. Only four capacitors and two air-wound inductors are needed. Maximum harmonic attenuation is achieved when the filter is built in a shielded box with two compartments. One section of the filter is placed in each compartment, as shown in **fig. 2**.

building the filter

A variety of aluminum boxes can be used for the filter. I suggest either the BUD "Minibox" or "Utility Cabinet." The size of the box depends upon the size of the components. For transmitters and receivers up to 100 watts output or so, I suggest 500-volt, dipped-mica capacitors, such as the Cornell-Dubilier CD series. Silver-mica capacitors of 1-kV "postage stamp" variety are also acceptable. For higher power levels, transmitting-type mica capacitors of 2.5 kV to 5 kV are suggested. (The voltage rating of the capacitor is not the limiting factor; it's the current-carrying ability. Roughly speaking, this is a function of the area of the capacitor plates. Physically small mica capacitors, while they might have a satisfactory voltage rating, usually cannot stand the rf current).

The inductors can be portions of miniature manufactured coils, such as the so-called "miniductors" made by several manufacturers. Or they may be self-supporting and made of heavy copper wire. The shield should be firmly bolted to all sides of the box. Coaxial receptacles are placed on the ends of the enclosure.

When completed, the filter can be adjusted by soldering a low-inductance ground strap to the lead between the two inductors where it passes through the hole in the shield. When this point is grounded, each coil and capacitor configuration may be dipped to resonance at the design frequency. All that is required is a slight adjustment of the coil turns. Once this is accomplished the box is sealed up.

using the harmonic filter

The filter is placed in the 50-ohm coaxial line between the antenna and station equipment. If an SWR meter is used, it should be placed between the filter and the transmitter. You don't want anything after the filter that might upset the filter's excellent harmonic rejection ability.

Remember, the filter rejects signals above its design frequency. It cannot be used on a higher frequency band. Essentially, it is a one band affair, and using it on a higher band will probably damage the capacitors.

the WØJZY mini-quad for 10 meters

The ham world abounds with mini-antennas and some of them work. Unfortunately most of them don't, so it is refreshing to find a mini-quad that delivers the goods. It was developed in 1978 by Leo Parra, WØJZY. I've heard him using this little antenna on 29.6 fm for several years, and his signal is outstanding. So I prevailed upon Leo to send me the information. Here's the story of his mini-quad antenna.

Electrically, the antenna is a loaded quad with a driven element and reflector (**fig. 3**). Four small toroids are placed in each loop, and these loading coils reduce the quad to about five feet on a side. Spacing between elements is 50 inches (127 cm).

The quad's driven element is fed at the center of the bottom wire with RG-58/U coaxial line. When fed directly, the radiation pattern is not quite symmetrical about the boom axis. The purist can place a balun at the feed point and re-establish pattern balance.

The front-to-back ratio is established with a short stub on the reflector element. A section of 300-ohm TV "ribbon" line is used, with the open ends shorted together. The stub is about 16 inches (40 cm) long to start with and it can be adjusted at the design frequency by listening to a local station off the back of the antenna and progressively shortening the stub by moving the short along the line toward the antenna. This can be done by gently seizing the stub with a pair of wire cutters and squeezing the cutters until they "nick" both wires of the line. If you inadvertently cut the line, it is no great loss and another one can be quickly substituted.

To get the antenna "in the ballpark," each element is dipped individually before assembly. The driven element is dipped about 400 kHz higher than the design frequency, and the reflector is dipped about 200 kHz lower than the design frequency. The loops are then assembled to the boom. For example, assume the antenna design frequency is 28.6 MHz. The driven element is dipped to 29 MHz (a one-turn loop placed across the loop's coaxial cable terminals will provide good coupling). The resonant frequency can be moved about by squeezing the turns equally on the loading coils or by adding or subtracting an inch or two of wire equally on each side of the quad loop.

The reflector is now dipped at 28.4 MHz. A small one-turn loop placed at the far end of the stub will do the job. Trim the stub length until you are on the nose. You can make a final check with a local signal on front-to-back ratio and you may not have to make any further adjustments.

Leo makes his adjustments with the loops in a vertical plane and with at least five feet clearance above ground to the bottom wire. A stepladder will do the job.

building the mini-quad

There are many ways of making the little quad. Leo uses a boom made of well-seasoned wood (Perhaps a closet pole? I didn't ask him what the wood was.) It is small enough in diameter to fit with a PVC pipe union, two of which are force-fitted onto the pole ends, (**fig. 4**). The union is for 3/4-inch water pipe. A short plug fits into the open end of the union and the quad arm supports are bolted to the plug. Each quad arm is made of a length of fiberglass rod, which looks suspiciously like a section of a fishing pole. They are very light and flexible. A banana jack is force-fitted on the end of each pole. The poles join at the center in an X-fitting (**fig. 5**) made of short sections of angle stock bolted together to form a hollow cross. The cross support, in turn, is bolted to the plug.

Leo's loops are made of insulated, flexible, test lead wire. Each quad loop is assembled with the loading coils and, at the points where the loop meets the fiberglass rods, the wire is broken and two large soldering lugs are soldered together and slipped over the banana jack's threaded shaft. A nut on the shaft is adjusted for tension. The wires are soldered to the lugs (**fig. 6**).

Before assembly, the whole mess has the rigidity of a wet noodle. But wait! Leo assembles the loops flat (on a picnic table, for example). He pushes the cross-shaped center spider onto each end of the boom, then pushes the fiberglass arms into the open ends of the spider. He then stretches the loops out, snaps the soldering lug assemblies over the ends of the banana jacks and tightens the tension bolts. With a little persuasion, the quad loop becomes amazingly rigid. The loops are then attached to the boom, the feedline (and balun, if used) attached to the driven loop, and the coax is brought back to the boom and taped along it.

When I assembled my WØJZY quad I used two eight-foot sections of a TV mast to support it about 15 feet (4.6 meters) in the air. The mast was held in a vertical position by placing it in a concrete base intended to hold a beach umbrella.

Sure enough, the little quad worked. Without any adjustment other

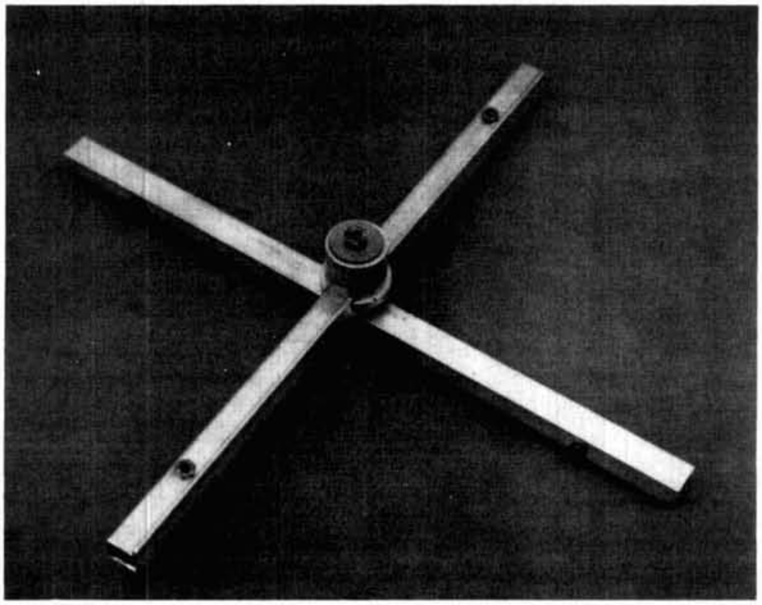


fig. 5. Quad cross-arm support is made of aluminum angle stock bolted together to form a square, hollow tube. Two tubes are mounted to plastic plug which slip fits within union attached to end of the boom. Fiber glass rods slip into ends of the X-fitting.

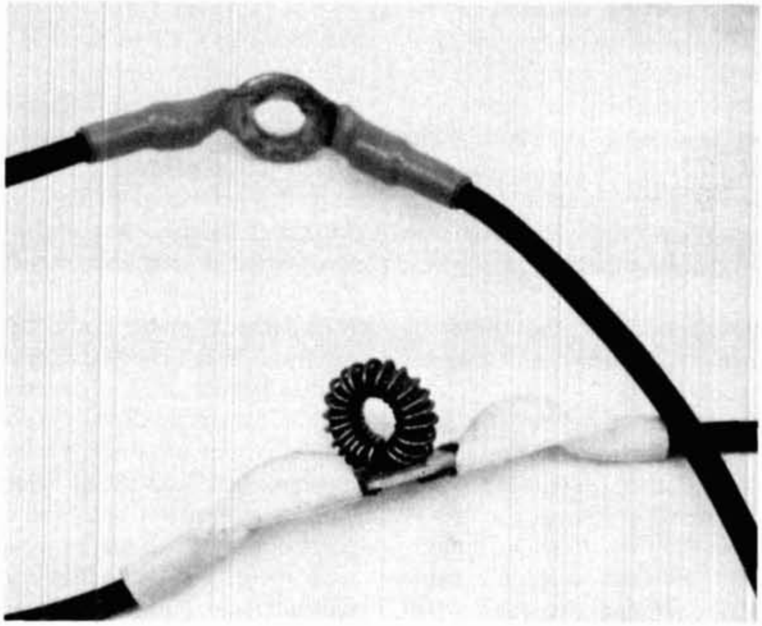


fig. 6. Close-up of the toroid coil and the double solder lugs. The powdered-iron coil is epoxied to a miniature insulator made of a scrap of printed circuit board. Antenna wires are soldered to ends of the board and heat-shrink tubing placed over the connections. Soldering lugs are sweated together to form a double-ended fixture which slips over the barrel of the inverted banana jack driven onto the tip end of the fiber glass support arm. Flexible, insulated test lead wire is used for element loops.

than dipping the elements, the quad exhibited about 10 dB front-to-back ratio at the design frequency. As for the gain — who can tell? Leo's signal with his mini-quad is excellent, and I worked many stations with this little antenna. In truth, it was not as good as my four-element Yagi atop a fifty-five-foot tower, but that is comparing apples and oranges. I'll tell you this — it was a *lot* better than a ground-plane antenna erected in the same place and at the same height.

I haven't given specific build-it-yourself dimensions but you should be able to construct your own mini-quad from this data. Leo had problems mounting the toroid coils in the antenna wire, but finally made up a small insulator out of dual-sided printed circuit board. He fastened the toroid to the board with epoxy cement and trimmed the copper face out at the center to form a simple insulated mounting. Each end of the winding, plus an antenna wire, was soldered to each end of the homemade insulator. I'm sure you can think of other means if you put your mind to it.

When properly made, the mini-quad can be erected or taken down in a few minutes. It makes an excellent antenna for Field Day, in addition to being an all-around good beam for those Amateurs who have restricted antenna space. No doubt the mini-quad can be modified for other ham bands. I'll leave that design modification up to you!

Note: For more information on quad antennas, refer to "All About Cubical Quad Antennas" by W6SAI and W2LX, available from Ham Radio's Bookstore.

ham radio

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More Details? CHECK—OFF Page 100

an armstrong beam rotator

In the early days of Amateur Radio, when beam antennas were somewhat of a novelty, the *Radio Handbook*[1] published a method showing how to turn the beam from inside your station. It was a good idea, but required a hole in the roof to accommodate the antenna mast. This requirement could pose all sorts of problems, depending on your roof structure and antenna-control position. (The device was directly beneath the antenna mast.) Other problems come to mind: what do you do if the rotor motor freezes after the warranty runs out? How do you seal the hole for the mast if you have a shingle roof?

Here's an alternative beam rotator that requires no electrical power, doesn't freeze, needs no long control cable, and needs no electrically driven direction indicator. All you need is some muscle power and a little ingenuity. And that's what ham radio is all about!

thrust bearing

Beam antennas that don't need a tower can be rotated by putting one end of the vertical support (pole or pipe) into a thrust bearing at ground level. (I chopped a round hole into my concrete patio, using a hammer and chisel.) You can also use ball bearings or something else more exotic. Or, a pipe flange slightly larger than the support pipe can be secured onto a greased platform mounted on the ground. Let your imagination run free!

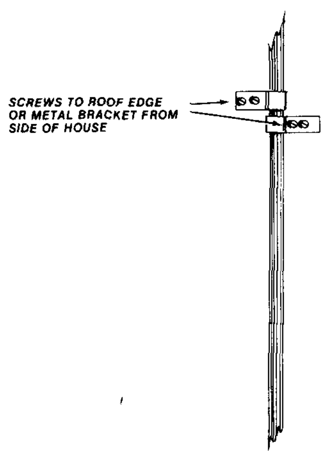

fig. 1. Antenna mast supports consist of pipe clamps secured to the edge of the roof or to a metal bracket mounted to the side of the house. Mast must be plumbed vertically and rotate freely inside the clamps.

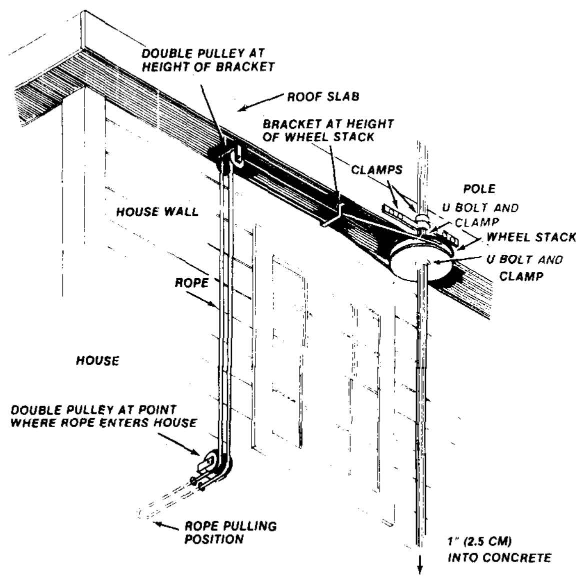

fig. 2. Suggested mounting arrangement for mast and drive wheel.

mast support

The thrust bearing support for the pole should be located by a plumb bob held from the next higher, or final, support for the pole, which is often at the edge of your roof. My roof consists of a flat concrete slab which extends a couple of feet further out than the house. If you have a Cape Cod or similar roof, you'll have to mount a bracket secured to the side of the house, with clamps attached to the bracket where the pole just misses the edge of your roof. One pipe clamp facing one direction and another clamp facing the other will do nicely (**fig. 1**).

Secure the pipe clamps with bolts or screws at the plumb-bob point so that the pole is held vertically but can be rotated in the clamps and thrust bearing a full 360 degrees.

Allow enough space between the pole and the side of your house to accommodate the radius of the

By John R. Schuler, KP4DM, Ruta Rural Buzon 2593A, Vega Baja, Puerto Rico 00763

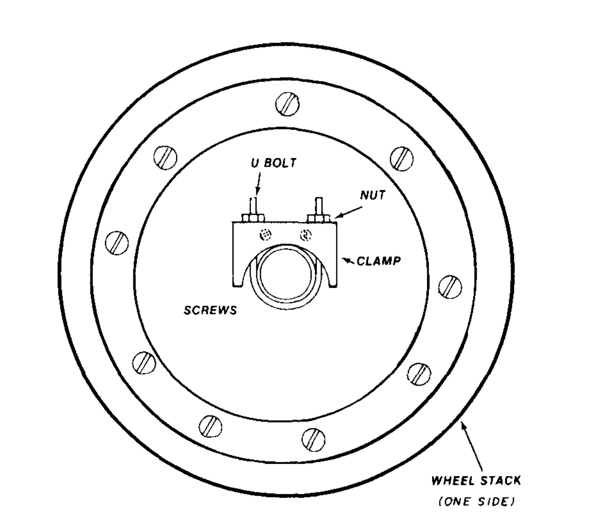

fig. 3. Detail showing method of securing wheel stack to the mast (one side only shown).

wheel-like device, which will allow you to rotate the antenna mast by means of a nylon rope and your arm muscles. Before you position either the thrust bearing or the pipe brackets with their supports, consider the radius of this wheel, keeping in mind the pole as the axis of the wheel. The wheel must be able to turn without scraping the side of your house. **Fig. 2** shows a suggested mounting scheme.

drive wheel

Recall the grooved wheel often found in dial-drive mechanisms. Cords are fastened to the wheel with springs in counter directions, so that pulling on one cord rotates the wheel one way, and pulling on the other cord rotates the wheel the other way. For my wheel, I used both ends of a wooden wire reel after removing the spindle between the ends. The hole in the center of each end piece is for supporting the pieces on a tube so that the wire may be unloaded by revolving the wheel. The hole should be big enough so that your pole mast can pass through it.

After the removal of the screws that hold the reel together, the spindle may be removed and discarded. You can probably find such a reel at a hardware store, the telephone company, power company, or an electrical contractor.

wheel assembly

Place both wheels over your pole and secure them together in several places with bolts and nuts. Holes for the bolts should be drilled about 1 inch (2.5 cm) in from the perimeter of the wheel. You now have a two-wheel stack. Next, get some sheet metal and cut pieces so that they may be secured with other bolts and nuts to the top and bottom of the two-wheel stack to form lips that extend the wheel perimeters about 2 inches. You have now made a channel for supporting the turning ropes so that they don't fall off the wheel (and are not pulled off when a little slack works into either rope).

The next step is to attach some U bolts and clamps to the top and bottom of the wheel stack with screws so that the wheel stack is held horizontally in place when you tighten the nuts of the U bolts (after placing stack over the pole.) See **fig. 3**.

Now decide where to place the double pulleys so that the end of the rope arrives inside the room at a convenient place. Add a few more feet of rope to allow yourself pulling room, a place to clamp the ropes when the beam points where you want, and a few more feet to cover the wheel stack perimeter.

dressing the pull rope

Purchase double this length of 1/4-inch nylon rope. Secure one end of this rope to the wheel stack perimeter and rotate the wheel more than 360 degrees either clockwise or counter clockwise. Pass the other end of the rope through the pulley sets, into and out of your shack, back through the pulley sets, and back to the wheel stack.

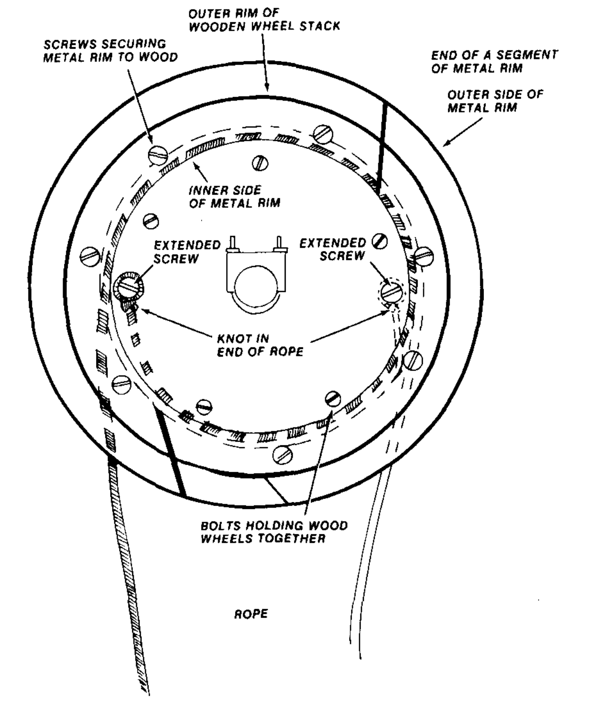

fig. 4. Method of dressing the pull rope around the wheel stack. A sheet metal extension around the periphery of the wheel provides a lip or trough for the pull rope.

Rotate the wheel stack in the opposite direction, keeping the rope you first secured taut, and place the end of the rope at a point again more than 360 degrees and again secure it near the perimeter. Now, by pulling on the first rope the wheel stack will rotate in one direction and by pulling on the other rope, the wheel stack will rotate in the other direction. The complete assembly of the wheel is shown in **fig. 4**. You are now ready to mount your Yagi to the pole and keeping the ropes taut, rotate your beam from the shack.

final touches

Final niceties include a) mounting a microphone holder to grip the rope in a taut position inside the station, b) painting the wheel stack, and c) marking the rope with compass directions. I applied paper labels marked with compass directions to the rope using Scotch™ tape so that, with everything taut, the marker at the bottom of the U-shaped rope facing the operator shows the direction in which the beam is pointed.

In my installation, a coat-hanger wire makes a support channel for the ropes where the distance between the pulleys and the wheel would otherwise allow the ropes to sag into a spot where they could snag the back door. It's best to lay out the whole arrangement in advance to avoid more pulleys and direction changes than are absolutely necessary. Naturally the more pulleys and direction changes, the more force needed to pull the beam to a new direction. Also don't forget that inertia and momentum play a part. It takes a strong pull maintained for a few seconds before the beam starts to turn and less to keep it turning. Additionally, ease up on the pulling as you near the direction you want, otherwise you will overshoot the direction.

At my house I have only two double pulleys and the coat hanger before the rope arrives at the wheel. Don't forget that you'll have to drill two holes through your window or house wall to allow the rope to enter and leave the pulling area. Keep the holes in line with the location of the last pulley to reduce frictional losses.

If you get a larger-diameter wheel stack, the rope will be easier to pull against frictional losses but a longer rope and a longer pull distance will be required because of the increased circumference. Also, the center hole may have to be much larger diameter than desired. A bicycle wheel rim could be used but might present more problems.

reference

1. *Radio Handbook*, Editors and Engineers Division, Howard W. Sams & Co., Inc., 1951 edition.

ham radio

short circuits

biquad bandpass filter

Author NØDE has written to point out that the two values of C in the biquad bandpass filter (June issue, **fig. 1**, page 70) should be equal for **eq. 1** to be correct. In **fig. 1** both values of C should be 0.1 μF. Also, placement of the left-hand IC on the PC board (**fig. 2**) is reversed; the notch should be oriented toward the lower edge of the circuit board.

updating the HW-2036

Complete circuit board layouts for WA4BZP's article (November, 1980, *ham radio*) are available by sending a stamped, self-addressed envelope to *ham radio*, Greenville, NH 03048.

quad variations

There is a drafting error in **fig. 4** of the ham notebook item "More Quad Variations" published in the October, 1980, issue of *ham radio*. The right-hand apex of the figure should be open; otherwise, the antenna will not resonate.

CW identifier

The following corrections should be incorporated in the article "Versatile CW Identifier," which appeared in the October, 1980, issue of *ham radio*: R2 is a 1-megohm pot. Pin 11 of U2 is listed twice; ignore the one that is grounded. R3 and C2 should be interchanged. In all cases, the PC board layout is correct as published.

antenna match

There are two errors in the article by Leonard H. Anderson that appeared in the January, 1981, issue of *ham radio*. In the right-hand column of page 59, the second line from the top of the page should read as follows: $= 153.04 \,/\, 3.9 = 39.24\ feet.$ **Equation 3**, in the same column, should read as follows:

$$\beta = arctan\ \frac{R_O(R_O - R_L)}{R_O X_L + R_L X} \qquad (3)$$

More Details? CHECK—OFF Page 100

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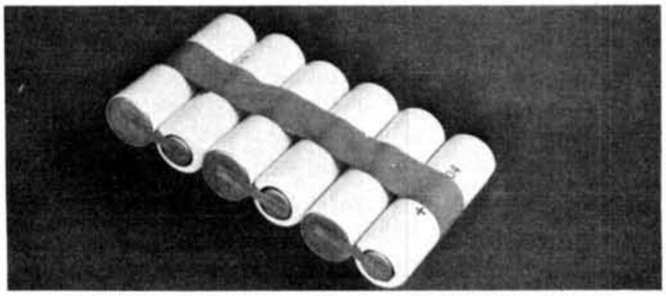

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F71-1115		connector with hex nut	10 1.99
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F71-1116		BNC male to female 90°	2.59
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6 pin plug	.85	UG 491	
F71-1121		BNC male to 80-239 female	3.00
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Din plug & jack		chassis mount	.80
5 pin male & female	.49	UG 914	
BNC UG260		BNC female to BNC female	
BNC male for RG 59U	1.25	RS-232 Hoods	1.00
BNC UG88U		RS-232 Male PCB type	2.00
BNC male for RG 58U		RS-232 Female PCB type	2.00
UG 273		Centronics male	6.99
BNC female to PL-259	2.50	F-59 connector for UG 59U	
UG 274		cable	100/13.95 or 10/2.00
BNC T			

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		MRF428	46.00	MRF629	3.45
		MRF428A	46.00	MRF648	33.35
		MRF426	15.50	MRF901	2.15
		MRF426A	15.50	MRF902	8.00
		MRF449	12.65	MRF904	3.00
		MRF449A	12.65	MRF911	3.00
		MRF450	13.80	MRF5176	3.00
				MRF8004	2.10
		MRF450A	13.80	BFR90	1.30
MRF203	P.O.R.	MRF452	15.00	BFR91	1.65
MRF216	31.00	MRF453/GE185	17.25	BFR96	2.20
MRF221	10.90	MRF454	19.90	BFW92A	1.15
MRF226	12.65	MRF454A	21.83	BFW92	1.00
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MRF247	34.00	MRF474	3.00	MMCM3960A	24.30
MRF262	9.20	MRF475	2.90	MWA120	10.00
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MRF412	25.30	MRF485	3.00	MWA220	10.00
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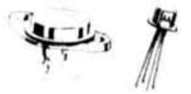
MRF422A	41.40	MRF502	1.04	MWA310	10.00

Transistors



		2N3959	3.85	2N5645	13.80
		2N3960JANTX	10.00	2N5842	8.00
		2N4072	1.80	2N5849	20.00
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		2N4429	7.00	2N5946	19.00
		2N4877	1.00	2N5862	57.50
		2N4959	2.30	2N6080	9.20
		2N4976	15.00	2N6081	10.35
2N2857	1.55	2N5070	18.40	2N6082	11.50
2N2857JAN	2.50	2N5071	20.70	2N6083	13.25
2N2949	3.60	2N5108	4.00	2N6084	15.00
2N2947	15.00	2N5109	1.70	2N6095	12.00
2N2950	4.60	2N5179	1.00	2N6096	15.50
2N3375	8.00	2N5583	4.00	2N6097	17.25
2N3553	1.57	2N5589	8.65	2N6166	40.25
2N3818	5.00	2N5590	10.35	2N6368	28.75
2N3866	1.30	2N5591	13.80	A210/MRF517	2.00
2N3866JAN	2.50	2N5635	10.35	BLY38	5.00
2N3866JANTX	4.00	2N5636	12.00	40280/2N4427	1.30
2N3925	10.00	2N5637	15.50	40281/2N3920	7.00
2N3948	2.00	2N5641	9.20	40282/2N3927	17.25
2N3950	25.00	2N5643	15.50	MMT74	1.04

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Plug Ins

Model#	Description	Make	Quantity	Price
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1A1	Dual Trace	Tektronix	4	$ 350.00 each
1A2	Dual Trace	Tektronix	1	$ 250.00 each
2A63	Differential Amp.	Tektronix	1	$ 100.00 each
1A4	Four Channel Amp.	Tektronix	4	$ 450.00 each
1S1	Sampling Unit	Tektronix	1	$ 250.00 each
3A72	Dual Trace Amp.	Tektronix	1	$ 250.00 each
53/54C	Dual Trace Calibrated Preamp	Tektronix	5	$ 79.99 each
3A75	Amplifier	Tektronix	1	$ 100.00 each
N	Sampling Unit	Tektronix	1	$ 100.00 each
1754A	Four Channel Amp.	Hewlett Packard	3	$ 90.00 each
1750B	Dual Trace Vertical Amp.	Hewlett Packard	1	$ 70.00 each
3T77	Sampling Sweep	Tektronix	1	$ 250.00 each
3T4	Programmable Sampling Sweep	Tektronix	1	$ 200.00 each
10A1	Differential Amp.	Tektronix	2	$ 100.00 each
3S76	Sampling Dual Trace	Tektronix	1	$ 250.00 each

Equipment

CDC5	Decade Capacitance	Cornell/Dubilier	4	$ 19.00 each
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204B	Dialamatic Volt Meter	Wavetek	1	$ 100.00 each
201	Dialamatic Volt Meter	Wavetek	1	$ 100.00 each
10411A	Horizontal Gain Calibrator	Hewlett Packard	1	$ 39.95 each
422	Oscilloscope	Tektronix	1	$ 900.00 each
PS163	Oscilloscope	Sencore	1	$ 409.95 each
175A	Oscilloscope Includes: plug ins; 1781B Delay Generator 1754A 4 Channel Amp.	Hewlett Packard	2	$ 400.00 each
71D	Capacitance/Inductance includes: 94A Digital Display 15A Bias Supply	Boonton Elect.	1	$ 200.00 each
2200	Filter	Krhon-Hite	1	
8100	Automatic Counter	Dana	1	$ 300.00 each
5245L	Electronic Counter includes: 5253A Freq. Counter 100-500 M.C.	Hewlett Packard	1	$ 950.00 each
6220	Frequency Multiple/Counter	Systron Donner	2	$ 150.00 each
1915A	Variable Transition Time Output	Hewlett Packard	1	$ POR
A202-292-4	Strip Chart Recorder	Gulton	1	$ 29.99 each
1784A	Strip Chart Recorder	Hewlett Packard	1	$ 150.00 each
414A	Auto Volt Meter	Hewlett Packard	1	$ POR
3200B	VHF Oscillator	Hewlett Packard	1	$ 750.00 each
431C	Power Meter	Hewlett Packard	1	$ 250.00 each
431B	Power Meter	Hewlett Packard	1	$ 150.00 each
TF1041C	Vacuum Tube Volt Meter	Marconi Instrument	1	$ 79.99 each
6328A/401B	Power Meter	Spectra Physics	1	$ 49.99 each
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75D	Direct Capacitance Bridge	Boonton Elect.	1	$1000.00 each
180A	Time Mark Generator	Tektronix	1	$ 49.00 each
1521B	Graphic Level Recorder	General Radio	1	$ 500.00 each
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HP3503	Microwave Switch .5-12.4 GHZ	Hewlett Packard		$ 50.00 each

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DX FORECASTER

Garth Stonehocker, K0RYW

last-minute forecast

Excellent conditions are forecast for the first and last weeks of February, with a dip of fair-to-good conditions in between. The excellent conditions should mean the higher frequency bands for long-haul DX can be counted on for some really good DX sessions. In between, in mid-month, the lower bands will be open for some all-night DX activity, if you're game. Some disturbed periods can be expected at that time about the 8th, 16th, and 25th. These periods should keep all of the bands interesting. Look for that intriguing new DX contact you haven't yet heard.

February propagation events

If you have fast mail service and live in New Zealand (South Island) or elsewhere in the South Pacific, the tip of South America, or the South Atlantic or Indian Ocean areas, you may be interested in this notice of a middle-size partial eclipse of the sun on January 25th from 0250 to 0634 UT. We will have to wait until June and July for any other eclipses this year. The moon will be full on the 8th.

No significant meteor showers appear in February. February, however, is often the month with the highest mean solar radio flux values this year. This is a result of the Earth's being closer to the sun during these winter months. The higher average radiation into the ionosphere from this 5 percent decrease in distance is significant. There is, however, an ionization build-up lag, so that February is often the high.

The winter months are also the most free from geomagnetic disturbances, which tend to lower the mid-latitude maximum usable frequencies (MUFs). Two factors are therefore working together to produce high MUFs in winter. In some years, though, the equinox periods, which provide favorable partial path alignment to the Earth, build high MUFs from solar flare activity at equinox. In these years the equinox periods might outdo the winter for the highest solar flux values. In either case the result is the highest daytime high frequency and VHF DX paths of the year. Geomagnetic activity pushes the ions toward the geomagnetic equator in Peru. The one-long-hop transequatorial propagation path results from refraction from the ionization maximums on each side of the geomagnetic equator and not reflecting from the Earth in between. Look for it in the late evening.

February is the month when changes in the ionosphere portend leaving the winter DX conditions of November, December, and January behind. The 10, 15, and 20 meter bands can be expected to open sooner in the morning and stay open longer into the evening. On the 40, 80, and 160 meter bands, which depend on darkness for their openings, the DX hours can be expected to shrink.

band-by-band summary

Six meters will open occasionally toward Europe — that is, to the east before noon, toward the south during noontime to afternoon, and toward the west and northwest in the late afternoon into early evening. The best openings are most likely transequatorial during high solar radio flux in evenings.

Ten and fifteen meters will exhibit the same pattern as 6 meters but will be open a longer part of the day. This is particularly true this month, since it is nearing springtime with its noticeably longer days and probably higher MUFs because of higher solar flux.

Twenty meters is a great band for everyone's pleasure, limited only by QRM. It should be open nearly every day and late into each evening to almost every part of the world. Best DX conditions can be expected just after sunrise and just before sunset for long skip.

Forty meters begins a transition into a nighttime band. Short skip during the daytime in winter, however, gives some interesting opportunities for working your close neighbors for the WAS certificate. Then, at evening, as the long skip (1000 to 2500 miles) develops, reach out for the far states and the WAC certificate. This band is very active to most areas of the world. In late afternoon the band will open to Europe, then swing around to South Africa and Central and South America, and then swing still farther into the Pacific by dawn.

Eighty and one-sixty meter DX conditions will be very good this month. Soon the atmospheric noise of the spring storms will give days of short skip QRN and local QRN. On toward summer the static will become bad enough that DX will have to be forgotten until fall. Take advantage of what's left of this year's quiet winter season. The directional pattern for these bands is similar to that of 40 meters. The low take-off angle of vertical antennas is very useful for DX here. Horizontal antennas are mainly short skip, high-take-off-angle radiators because of being so close to the ground. Look for particularly interesting DX as these bands come in (open) near sunset and go out at sunrise.

ham radio

FEBRUARY

WESTERN USA

	2300	2200	2100	2000	1900	1800	1700	1600	1500	1400	1300	1200	1100	1000	0900	0800	0700	0600	0500	0400	0300	0200	0100	0000	GMT
FEBRUARY	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	PST
ASIA FAR EAST	10	15	15	20	20	20	20	20	—	—	—	—	—	—	—	—	20	20	15	15	10	10	10	10	N
EUROPE	20	20	20	20	15	10	10	10	15	—	—	—	20	20	20	20	40	40	40	40	—	—	—	—	NE
S. AFRICA	10	10	10	10	10	10	10	10	10	15	—	—	—	—	—	20	20	20	20	20	20	15	15	15	E
S. AMERICA	10	10	10	10	10	10	10	10	20	40	40	20	20	20	20	20	20	20	20	15	15	10	10	10	SE
ANTARCTICA	—	—	—	—	—	—	—	—	80	40	40	40	40	20	20	20	20	15	15	15	15	15	15	15	S
NEW ZEALAND	10	10	10	10	10	10	15	15	20	40	40	40	40	20	20	20	20	15	15	10	10	10	10	10	SW
OCEANIA AUSTRALIA	10	10	10	10	15	15	20	20	20	40	40	40	20	20	20	20	20	15	15	15	10	10	10	10	W
JAPAN	10	10	15	—	—	20	20	20	40	40	40	40	20	40	40	20	20	20	15	15	15	10	10	10	NW

MID USA

	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	MST
ASIA FAR EAST	10	15	20	20	20	20	20	—	—	—	—	—	—	—	—	—	20	20	15	15	10	10	10	10	N
EUROPE	20	20	20	15	15	15	10	10	10	15	20	—	—	—	—	—	40	40	40	40	40	20	20	20	NE
S. AFRICA	15	15	10	10	10	10	10	10	10	20	—	—	—	—	—	—	—	20	20	20	20	20	15	15	E
S. AMERICA	10	10	10	10	10	10	10	10	10	15	20	40	40	20	20	20	20	20	20	20*	15	15	10	10	SE
ANTARCTICA	15	—	—	—	—	—	—	—	—	—	40	40	40	40	20	20	20	20	20	20	20	15	15	15	S
NEW ZEALAND	10	10	10	10	10	10	15	15	20	20	20	20	20	20	20	20	15	15	15	15*	10	10	10	10	SW
OCEANIA AUSTRALIA	10	10	10	10	15	15	15	15	20	20	40	40	40	40	40	20	20	20	20	15	10	10	10	10	W
JAPAN	10	10	20	—	—	—	20	20	20	—	—	—	—	—	20	20	20	20	20	15	15	10	10	10	NW
	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	CST

EASTERN USA

	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	6:00	5:00	4:00	3:00	2:00	1:00	12:00	11:00	10:00	9:00	8:00	7:00	EST
ASIA FAR EAST	20*	20	—	—	—	—	—	20	20	20	20	—	—	—	—	—	—	—	—	—	20	15	15	15	N
EUROPE	20	20	15	15	15	15	10	10	10	10	15	15	15	20	40	40	40	40	40	40	40	20	20	20	NE
S. AFRICA	15	15	15	10	10	10	10	10	10	10	10	10	—	—	—	—	—	20	20	20	20	20	15	15	E
CARIBBEAN S. AMERICA	10	10	10	10	10	10	10	10	10	10	10	10	20	40	40	20	20	20	15	15	15	15	10	10	SE
ANTARCTICA	15	15	15	—	—	—	—	—	—	—	—	—	40	40	20	20	20	20	20	20	20	20	20	15	S
NEW ZEALAND	10	10	10	10	10	10	15	15	15	20	20	80*	80*	40	20	20	20	20	15	15	15	15	10	10	SW
OCEANIA AUSTRALIA	10	10	10	10	—	—	20	20	15	15	20	20	40	40	20	20	20	20	—	—	15	15	10	10	W
JAPAN	15	10	20	20	—	—	—	—	20	20	—	—	—	20	20	20	20	20	20	20	20	20*	15	15	NW

*Look at next **higher** band for possible openings.

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CADILLAC OF QSL CARDS, 3 to 4 colors, send $1 for samples (Refundable). Mac's Shack, P.O. Box 43175, Seven Points, TX 75143.

FOR SALE: Two LM 54 self-supporting elect. crank tilt-over towers, Collins 32S3, 75S3B, 72S1, KWM2, 30L1, Heathkit HW 100, Icom 2m FM 280, 22S, 230, 221 (all mode) Cobra 220 FM, Lunar amp, klitzing 10 in 100 out Yaesu FT 625 RD Alda 103 transceiver B & H scope 5 MHz bandwidth 2 Ham ann. and controls, elevation rotor KR 500 Hy-Gain Tribander 4 F9Ft. ant. No reasonable offer refused. Please call or write: Glenda Brault, 2412 E. Valley Oaks Dr., Visalia, CA 93291. (209) 734-0884.

SELL: Transmitting tube collection, cheap! Call weekends only. 503-773-5970.

WANTED: RG-17/u connectors, UG-154/u plugs and UG-352/u receptacles. RG-17a/u coax cable and 7/8" 50 Ohm hardline or large diameter Heliax. Also, good condition CDE Tailtwister rotator. Milton R. Baronovic, 10073 McCartney, St. Louis, MO 63137. (314) 867-5849 evenings. WAØVIO.

MOBILE OPERATORS: Anteck's Mobile Antennas cover 3.2 to 30 MHz inclusive, with no coil changing. 50 Ohms input. Two models, the MT-1 MANUAL, MT-1RT REMOTE-TUNED from the operators position. Uses two Hyd. Pumps and Motors. MT-1 $129.95, MT-1RT $240.00 plus UPS postage. Check your local dealer or write for Dealer List and Brochure. ANTECK, INC., Route One, Box 415, Hansen, ID 83334. 208-423-4100.

CODE SIMPLIFIED. Powerful algorithmic design reduces Morse to logic-compatible serial, parallel, ASCII. Just $169. includes shipping. Telecraft Laboratories, Box 1185, E. Dennis, Mass. 02641.

NOTICE: I will pay $1/oz. for gold-plated pins, connectors, transistor leads and PC board fingers. W6TWT, Ron Guard, 10105 Stern Ave., Cupertino, CA 95014. (408) 252-5067.

HAM RADIO FANATICS! You need the W5YI Report — Twice monthly award-winning insider newsletter. 24 issues — $18.00. Sample issue SASE (2 stamps). W5YI, Box #10101-H, Dallas, Texas 75207.

ATLAS DD6-C Digital Dial $120.00 plus $4.00 UPS. NEW, while they last. Mical Devices, P.O. Box 343, Vista, CA 92083.

FOR SALE AS LOT ONLY: TS-830S, MC50 mike, MFJ 941c tuner, mini quad beam and balun. AR40 rotator + cables. As separate: Eico 720 CW Xmitter, Heath VF-1 VFO and surplus BC-221R freq. meter. All best reasonable offer, FOB NY. K2GFN.

RTTY JOURNAL-EXCLUSIVELY AMATEUR RADIOTELETYPE, one year subscription $7.00. Beginners RTTY Handbook $5.00, RTTY Index $1.50. P.O. Box RY, Cardiff, CA 92007.

DRAKE R7/DR7, MS7, NB7A. 0.3, 4.0, 6.0 accessory filters, service manual. Mint condition. $1400. obo. Barry Megdal (213) 269-3525 or (213) 792-3726. 13559 Haynes St., Van Nuys, CA 91401.

RECIPROCATING DETECTOR Handbook, $10.50 ppd. Peters Publications, 19 Loretta Rd., Waltham, Mass. 02154.

JOHNSON OR COLLINS WANTED. Valiant, 500, or KWS-1. Must be clean, perfect on CW. Also need 75A4 500 Hz filter. Pickup distance on rigs. K1IFJ, Box 502, Ridgefield, CT 06877. (203) 438-5944.

CUSTOM EMBROIDERED EMBLEMS — Your design, low minimum. Informational booklet. Emblems, Dept 65, Littleton, New Hampshire 03561.

WANTED: Pre-1950 TV sets and old TV Guides. Jeff Kadet, W3CRH, Box 90-HR, Rockville, MD 20850. (301) 654-1876.

MANUALS for most ham gear made 1937/1970. Send $1.00 for 18 page "Manual List", postpaid. HI-MANUALS, Box H802, Council Bluffs, Iowa 51502.

SWAP — Fanon 6-HLU Scanner, 30-50, 146-175, 450-512 MHz for general coverage receiver. K2VGV. (212) 461-5933.

SATELLITE TELEVISION: Information on building or buying your earth station. Six pages of what's needed, where to get it, costs, etc. $4.00 to Satellite Television, RD #3, Oxford, NY 13830. Build your own parabolic antenna. Book also available. Send SASE for details.

SB-102, SB-600, PS, CW filter. MINT honest. Original owner. K5PA, 12412 Mossybark, Austin, TX 78750. (512) 258-4959. $390.

NOVICE SHACK STARTER! Heath HR-1680 Rcvr — $150; HS-1661 Spkr — $15; SBA-104-1 Noise Blanker — $25. Package price — $175. Knight KG-620 VTVM — $20;

More Details? CHECK—OFF Page 100

Heath IG-102 RF Signal Generator — $25. Both for $40. Heath HM-102 SWR/wattmeter — $30. I ship it all for $240! George Ritter, WB8EPE, (614) 587-7290 (work); (614) 366-3970 (home).

HAM RADIO MAGAZINE issues 1968 through 1975 (excl. April 1968) in binders. Reasonable offer. Henry Ingwersen, RFD 1, Charlotte, VT 05445. (802) 425-2096.

SOLAR PANELS: Surplus, big savings!! Several sizes to choose from. Ideal for repeater. Write or call for complete info. Possible trade? Bob (K6CKD) 714-581-6369, 21946 Yellowstone Ln., El Toro, Calif. 92630.

FOR SALE: Drake DC-4 mobile power supply $49. Heathkit sixer $20. KØGXI. 4018 Ontario, Ames, IA 50010.

WANTED: Help in completing the largest collection of Hallicrafter equipment in the world. Urgently needed are receivers with aluminum colored panels, back lighted plastic dials with "airplane" hands, early transmitters, unusual accessories, etc. Chuck Dachis, WD5EOG, "The Hallicrafter Collector," 4500 Russell Drive, Austin, Texas 78745.

DATABOOKS, IC, TTL, electronics books and engineering magazines from Eng. library. Exc. modern material. Also some older Ham magazines & service manuals. Send SASE for 2-pg. list. J.B. Cohen, 200 Woodside Ave., Winthrop, MA 02152.

ELECTRONIC BARGAINS, CLOSEOUTS, SURPLUS! Parts, equipment, stereo, industrial, educational. Amazing values! Fascinating items unavailable in stores or catalogs anywhere. Unusual FREE catalog. ETCO-012, Box 762, Plattsburgh, NY 12901. SURPLUS WANTED.

WANTED: Racal receiver and Racal accessories or similar to RA6217B or E, MA 350B and MA6314 synthesizers, RA6377 comparator, MA6355 oscillator, MA210 Racalator. Equipment can be incomplete or with electrical fault. Need Racal books. John, KB7HC, Box 106, APO NY 09241.

WANTED: Two way service test gear. Need signal generator and mod. monitor. Please state your price, what you have, what freq. range, condition, etc. WA4MIT OK in C.B. (205) 339-9677.

FT-101E, Fan, SSB CW filters installed. Includes manual, mic, new tubes. Excellent condition. $550. plus shipping. John Skubick, K8JS, 791-106 Ave., Naples, FL 33940. (813) 597-3641.

HAM RADIO REPAIR, experienced, reasonable, commercial licensed. Robert Hall Electronics, P.O. Box 8363, San Francisco, CA 94128. W6BSH, (408) 292-6000.

GENERAL RADIO 1210-C OSCILLATOR, $10. 1551-B sound meter, $65. 1217-A pulser, $15. H-P 200C oscillator, exc. $25. 400H-R VTVM, exc. $28. 340-A Noise-Fig. meter, $55. 212A pulse gen. $15. 412A f/parts, $20. Dana 5600 DVM, $45. 20A Variac w/case, exc. $45. Ballantine VTVM's 310A, $20. 302B, $6. National CRU oscilloscope, $45. Keithley 414 micromicroammeter, $15. RCA WV-84C microammeter, $6. Measurements 65-B, repairable, $75. Regatron programmable power supply, $10. Lacing cord, mil-spec, $3/spool. J.B. Cohen, 200 Woodside Ave., Winthrop, MA 02152. (617) 846-6312.

RUBBER STAMPS: *3 lines $3.25 PPD. Send check or MO* to G.L. Pierce, 5521 Birkdale Way, San Diego, CA 92117. SASE brings information.

WANTED: Digital dial frequency counter for TS820 Kenwood. Please write price to Dave, PB 76, Fenwick, LØS ICØ, Ontario, Canada.

MOBILE IGNITION SHIELDING provides more range with no noise. Available most engines. Many other suppression accessories. Literature, Estes Engineering, 930 Marine Dr., Port Angeles, WA 98362.

KENWOOD TS180S w/DFC, cw/dual SSB filters, $650.00; Drake R4B, 10 extra crystals, $335; Sony ICF-2001 digital AM/FM/SW portable, $255; Millen 90651 G.D.O., coils w/rack, probe, $45.00; MFJ-752 SSB/cw superfilter, $40.00. All fine, w/manuals. WA7ZYQ, (208) 245-2070.

QSL ECONOMY: 1000 for $12. SASE for samples. W4TG, Box F, Gray, GA 31032.

MUST SELL: Radios, assembled boards, parts from pilot line build of the high quality Amtech 225A synthesized 2m FM transceiver. Business size SASE (40 cents postage) for list. Very low prices! Bill Weaverling, 3060 27th Ave., Marion, Iowa 52302. (319) 377-7927.

TUBES, TUBES wanted for cash or trade: 340TL, 4CX1000, SCX1500. Any high power or special purpose tubes of Eimac/Varian. DCO, 10 Schuyler Avenue, No. Arlington, NJ 07032. (800) 526-1270.

PICTURE QSLs — quality photo and standard cards. Free samples and information. Planet Publishing, P.O. Box B, Martinsville, IL 62442.

MATCHING NETWORKS AND FILTER DESIGN are fast and easy with your HP-41C and our programs. Write Software Specialties, Inc., Dept. HR/1, Box 329, Springboro, OH 45066.

WANTED: MS connectors (new or used), synchros, etc. Send list, Bill Williams, P.O. 7057, Norfolk, Virginia 23509.

WANTED: Micor and Master II base stations, 406-420 MHz. Any solid state 2 and 6 GHz microwave equipment, AK7B, 4 Ajax Place, Berkeley, CA 94708.

MAKE HAM RADIO FUN! Supplement your learning programs with a motivational hypnosis cassette. Tape #3, Learning the Code; Tape #4, Breaking the Speed Barrier; Tape #7, Electronic Theory. Free catalog. For tapes, $10.95 each to Gem Publishing, 3306 North 6th St., Coeur d-Alene, ID 83814.

VERY in-ter-est-ing! Next 5 issues $2. Ham Trader "Yellow Sheets", POB356, Wheaton, IL 60187.

EXCELLENT OPPORTUNITY IN SUNNY SOUTHWEST. Join our staff of 20+ technicians. Motorola MSS servicing 2-way radio, mobile telephones, micro-wave and CCTV equipment. Openings at all levels for technicians. Contact: Jim Strickland, K5EXB, Circle Communications Co., 2740 South Freeway, Ft. Worth, Texas 76104. Phone: (817) 923-7334.

CB TO 10 METER PROFESSIONALS: Your rig or buy ours — AM/SSB/CW. Certified Communications, 4138 So. Ferris, Fremont, Michigan 49412; (616) 924-4561.

AFC SEMI-KITS! Stop VFO drift. See June 1979 HR. $55.00 plus $3.00 UPS. Mical Devices, P.O. Box 343, Vista, CA 92083.

HAMS FOR CHRIST — Reach other Hams with a Gospel Tract sure to please. Clyde Stanfield, WA6HEG, 1570 N. Albright, Upland, CA 91786.

BUY-SELL-TRADE Send $1.00 for catalog. Give name address and call letters. Complete stock of major brands new and reconditioned amateur radio equipment. Call for best deals. We buy Collins, Drake, Swan, etc. Associated Radio, 8012 Conser, Overland Park, KS 66204. (913) 381-5900.

SATELLITE TELEVISION...HOWARD/COLEMAN boards to build your own receiver. For more information write: Robert Coleman, Rt. 3, Box 58-AHR, Travelers Rest, SC 29690.

Coming Events

ACTIVITIES

"Places to go..."

FLORIDA: The Treasure Coast Hamfest, February 20, Vero Beach Community Center. Prizes, drawings. Q.C.W.A. luncheon. Admission $2.00 advance; $2.50 door. Talk in on 146.13/73, 146.04/64, 146.52/52, 222.34/223.94. For information: P.O. Box 3088, Beach Station, Vero Beach, Florida 32960.

PLAYBOY CLUB: Plan ahead now to attend the ARRL Hudson Division Convention, October 30-31, 1982, at the Playboy Club, Great Gorge, McAfee, NJ. For info send SASE to HARC, Box 528, Englewood, NJ 07631.

INDIANA: *The LaPorte Amateur Radio Club's Winter* Hamfest, Sunday, February 28, Civic Auditorium, 8 AM. Donation: $2.50 door. Reserved tables $2.00 each. Write: P.O. Box 30, LaPorte, IN 46350.

MICHIGAN: The 12th annual Livonia Amateur Radio Club's Swap 'n Shop, Sunday, February 28, 8 AM to 4 PM, Churchill High School, Livonia. Door prizes, refreshments, free parking, tables. Talk in on 146.52 simplex. Reserved table space available. For information SASE to: Neil Coffin, WA8GWL, Livonia ARC, P.O. Box 2111, Livonia, MI 48151.

IOWA: The Davenport Radio Amateur Club's 11th annual Hamfest, Sunday, February 28, Davenport Masonic Temple, 7th and Brady Streets. Tickets $2 advance; $3 door. Tables $5 each plus $2 electrical hookup charge. 8 AM to 4 PM. Talk-in on WØBXR/Rpt at 146.28/.88 MHz. For tickets/tables write: Dave Johannsen, WBØFBP, 2131 Myrtle, Davenport, Iowa 52804.

TENNESSEE: See World's Fair while attending 1982 Knoxville Hamfest and ARRL Delta Division Convention, Memorial Day Weekend (May 22-23). DX, computer, and technical forums: air-conditioned exhibit area; large indoor/outdoor flea market. Make this Tennessee's largest hamfest. For more information: (dealers, tickets, reservations) N4BAQ, 5833 Clinton Hwy., Suite 203, Knoxville, Tenn. 37912.

More Details? CHECK—OFF Page 100

the ham notebook

CW BFO crystal for the 75S-3

The Collins 75S-3 receiver uses the upper sideband BFO crystal (456.350 kHz) for reception of both upper sideband and CW. The EMISSION switch selects either of two filters, but the BFO crystal frequency remains the same. This provides a beat note of 1350 Hz when the signal is properly positioned within the filter passband. I found the note to be too high for my taste and got tired of using the tunable BFO each time for CW reception.

Investigation showed that EMISSION-switch section S9 has two lugs jumpered where the upper sideband BFO crystal (Y16) is attached, although the schematic does not show this. By removing the jumper, an extra CW BFO crystal was added, which creates a more pleasing beat note when using the CW filter position. In my case, a 455.300-kHz crystal (HC-33/U holder) was used, which produces a 300-Hz beat note.* See **fig. 1**.

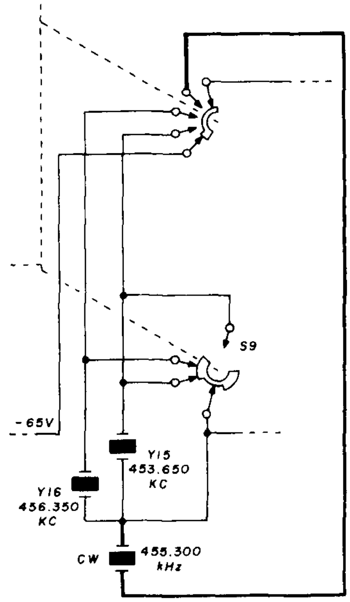

fig. 1. Heavy line shows addition of new CW BFO crystal to Collins 75S-3.

After cutting the jumper between the two switch lugs, ensure that the upper sideband crystal is soldered to its proper lug, then add the new CW crystal between its lug and the crystal common point at the terminal strip immediately behind the switch. Both lugs are easily accessible. With the receiver upside-down, front-panel toward you, the two lugs used are at the 11 o'clock upper sideband positions on the rear-most wafer.

When calibrating the receiver for operation close to the band edges, caution should be observed because the beat will differ between the upper sideband and CW positions by the frequency difference between the two crystals. This is the same precaution that must be taken when the tunable BFO is used. Once calibrated in a particular position, the setting will be accurate for that position of the EMISSION switch.

I've found, as have many others, that the lower CW beat note is much less tiring to listen to. Most of the high-frequency hiss and noise are eliminated, and copy approaches that of listening to a monitoring signal. The switch-selected CW position now provides both filtering and the preferred beat note at the flick of the switch.

Paul K. Pagel, N1FB

*F_{xtal} minus i-f equals F_{beat}; that is, *455.300 − 455.000 = 0.3, or 300 Hz.*

an rf power divider

Here's a variable rf power divider that's particularly suitable for UHF applications, and which:

1. Presents a matched load to the power source at all levels of power division.

2. Maintains a constant (90-degree) phase relationship between the two outputs.

3. Can be adjusted with a single control.

Consider the circuit of **fig. 2**. Matched loads are assumed at outputs **A** and **B**. If all components are normalized to the characteristic impedance of the lines, the matched loads will be unity, and the two shunt susceptances represented by the dual sliding shorts will be characterized as jB. Thus, at **A** we'll have an admittance of $1 + jB$. At point **C** there will be merely the susceptance jB, which when translated a quarter wavelength to **B** and combined with the load, becomes $1 + \frac{1}{jB}$. Both these complex admittances will again be inverted as they are translated an additional quarter wavelength to be combined at the input port. The input admittance thus becomes:

$$\frac{1}{1 + jB} + \frac{1}{1 + \frac{1}{jB}}, \text{ which becomes}$$

$$\frac{1}{1 + jB} + \frac{jB}{1 + jB} = 1 \qquad (1)$$

Therefore, as long as the loads are matched, and the two shunt susceptances are equal, a matched load will be presented to the power source.

Obviously the sum of the two output powers must equal the power input to the divider. Power, of course, is proportional to the square of the voltage, and as $sin^2\theta + cos^2\theta = 1$, the amplitudes of the output voltages must have a sine-cosine relationship.

A vector diagram, showing the action as the length of the sliding shorts vary from zero to a quarter wavelength, is shown in **fig. 3**. With

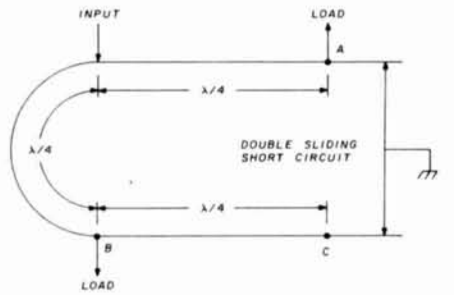

fig. 2. Power divider schematic. The double sliding short provides controlled shunt susceptances to obtain unity input admittance and thus a matched load to the input at all levels of power division.

a length of zero, output **A** (**fig. 2**) is shorted, while the quarter-wavelength, **BC**, allows output **B** to receive the full output.

When the lengths are increased to a quarter wavelength, the picture is reversed. Intermediate points will lie

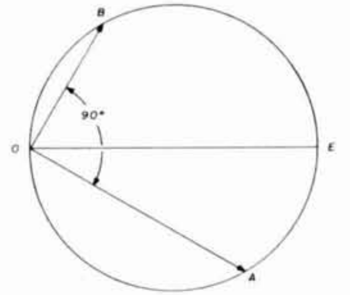

fig. 3. Vector diagram illustrating what happens in the power divider as the length of the sliding short circuit varies from zero to a quarter wavelength.

on semicircles, with **OE** as a common diameter. The geometry of the figure further dictates that the phase angle between **OA** and **OB** is 90 degrees at all times.

A dual sliding short is indicated in **fig. 2**. However, any method of producing identical controlled shunt susceptances will give the same results. If a zero-to-total changeover is not required, a small, two-section variable capacitor may be the answer. In a stripline configuration, a dielectric element can be used to vary the electrical length of the lines at **A** and **C**.

Henry S. Keen, W5TRS

KENTUCKY: The annual Glasgow Swapfest, Saturday, February 27, 8 AM til everyone goes home, Glasgow Flea Market Building, 2 miles south of Glasgow off Highway 31E. Large heated building, free parking, door prizes, free coffee, large flea market and a friendly gathering. Admission: $2.00 pp. One free table per exhibitor with extras available at $3.00 each. Talk-in on 146.34/94 (primary) or 147.63/03 (alternate). Additional information: WA4JZO, 121 Adairland Ct., Glasgow, KY 42141.

MICHIGAN: The Cherryland Amateur Radio Club's ninth annual Swap 'N Shop, Saturday, February 13, Immaculate Conception Middle School gymnasium, 218 Vine Street, Traverse City. Doors open 8 AM to 2:30 PM. General admission $2.50, single tables $3.00. Talk-in on 146.85 or 146.52 simplex. For information: Jerry Cermak, K8YVU, Chairman, 3905 Slusher Rd., Traverse City, MI 49684. Please SASE.

NEW JERSEY: The Split Rock Amateur Radio Association's annual equipment auction, Thursday, February 25, Morris Plains VFW Post #3401, Route 53, Morris Plains. Doors open 7 PM. Auction 8 PM sharp. Admission free. Please limit items to working electronic equipment — no junk. Bag or box loose parts. The club will take 10% commission on all individual sales up to $50. Over $50, a $5 commission will be taken on individual sales. Payable in cash only. Refreshments and plenty of parking. Rain date: Thursday, March 4, same time/location. For information: P.O. Box 3, Whippany, New Jersey 07981.

NEW YORK: Oneida County Mini-Fest and Flea Market, February 20, Oneida County Airport, Oriskany. For further details: (315) 736-0184.

OHIO: The Mid-Winter Hamfest/Auction, Sunday, February 14, Richland County Fairgrounds, Mansfield. Doors open 8 AM. Tickets: $2.00 advance; $3.00 door. Tables $5.00 advance; $6.00 door. Half tables available. Talk-in on 146.34/94. For info, advance tickets or tables, SASE to: Harry Frietchen, K8HF, 120 Homewood Road, Mansfield, OH 44906. (419) 529-2801.

OHIO: The Cuyahoga Falls Amateur Radio Club's 28th annual electronic equipment Auction and Flea Market, Sunday, February 28, North High School, Akron. 8:30 AM to 4 PM. Tickets $2.00 advance and $2.50 door. Prizes: First — Kenwood TS130S, second — Icom 3AT, third — Icom 2AT. A 16k TRS-80 Model III will be raffled at $2 a chance. Sellers bring own tables or rent for $2. Free parking. Check-in on 146.04/64. Details from CFARC, P.O. Box 6, Cuyahoga Falls, OH 44222 or call K8JSL (216) 923-3830.

VIRGINIA: The Vienna Wireless Society's 9th annual WINTERFEST '82, February 28, 8 AM, Community Center, 120 Cherry Street, Vienna. Displays, dealers, indoor flea market and "Frostbite" tailgating. Tables $5 and $10. Prizes include a Kenwood TS-830S hf transceiver, Icom IC-25A, mobile 2m rig, Santec HT-1200 handheld. Hourly drawings for accessories and books. Tickets $3.00 includes one chance for prizes. Talk-in on 31/91 and 146.52 simplex. For info: SASE to Winterfest '82, Vienna Wireless Society, P.O. Box 418, Vienna, Virginia 22180 or call Ray Johnson (703) 938-8313.

CINCINNATI ARRL '82: Hamilton County ARPSC invites all hams to participate in the second annual Ohio State Convention. Two full days of amateur activities; forums, meetings, exhibits, flea market, and more! This ALL INDOORS activity will take place on Saturday and Sunday, February 27 & 28. For further information contact Cincinnati ARRL '82, Committee for Amateur Radio, P.O. Box 46311, Cincinnati, OH 45246. Dealer and exhibitor inquiries invited. Registration $4, Flea Market $3.

NEW JERSEY: The Old Bridge Radio Association's second annual auction, February 28, Cheesequake Firehouse, Route 34 (Perrine Road), Old Bridge. Computer, electronic and Amateur Radio equipment. Admission: $1.00 includes chance at many door prizes. Doors open 11 AM. Sale starts 11:30. Club commission — 10% on first $100 of selling price; 5% on remainder. Refreshments available. Talk-in on .72/.12, .34/.94 and .52. For information: Fred, WA2BJZ. (201) 257-8753.

OPERATING EVENTS

"Things to do..."

FEBRUARY 7 THROUGH 9: NH-VT QSO Party. 2100A, 7 Feb. to 0500Z, 8 Feb.; 1100Z, 8 Feb. to 0100Z, 9 Feb. Frequencies: Phone — 3930, 3960, 7230, 7260, 14280, 14320, 21360, 28570, 50110, 144.2. CW — 3530, 3760, 7030, 7130, 14080, 21060, 21150, 28070, 144.1. Exchange: NH, VT stations send QSO number, RST, County, State. Others send QSO number, RST, State, Country. Certificate awards. SASE for results. Send logs/facsimiles, name, license class, address, NLT 15 March 1982 to: Rex Lint, K1HI, 10 Hartwood Dr., Merrimack, NH 03054.

Continued on page 83

HAM CALENDAR

February

W1AW Schedule

October 25, 1981-April 24, 1982 MTWThFSSn = Days of Week Dy = Daily
W1AW code practice and bulletin transmissions are sent on the following schedule

EST	Slow Code Practice	MWF: 9 A.M., 7 P.M., TThSSn: 4 P.M., 10 P.M.
	Fast Code Practice	MWF: 4 P.M., 10 P.M., TTh: 9 A.M., TThSSn: 7 P.M.
	CW Bulletins	Dy: 5 P.M., 8 P.M., 11 P.M., MTWThF: 10 A.M.
	RTTY Bulletins	Dy: 6 P.M., 9 P.M., 12 P.M., MTWThF: 11 A.M.
	Voice Bulletins	Dy: 9:30 P.M., 12:30 A.M.
CST	Slow Code Practice	MWF: 8 A.M., 6 P.M.; TThSSn: 3 P.M., 9 P.M.
	Fast Code Practice	MWF: 3 P.M., 9 P.M.; TTh: 8 A.M.; TThSSn: 6 P.M.
	CW Bulletins	Dy: 4 P.M., 7 P.M., 10 P.M.; MTWThF: 9 A.M.
	RTTY Bulletins	Dy: 5 P.M., 8 P.M., 11 P.M.; MTWThF: 10 A.M.
	Voice Bulletins	Dy: 8:30 P.M., 11:30 P.M.
PST	Slow Code Practice	MWF: 6 A.M., 4 P.M., TThSSn: 1 P.M., 7 P.M.
	Fast Code Practice	MWF: 1 P.M., 7 P.M., TTh: 6 A.M., TThSSn: 4 P.M.
	CW Bulletins	Dy: 2 P.M., 5 P.M., 8 P.M., MTWThF: 7 A.M.
	RTTY Bulletins	Dy: 3 P.M., 6 P.M., 9 P.M., MTWThF: 8 A.M.
	Voice Bulletins	Dy: 6:30 P.M., 9:30 P.M.

SUNDAY	MONDAY	TUESDAY	WEDNESDAY	THURSDAY	FRIDAY	SATURDAY
	WEST COAST BULLETIN – 9PM PDT (8PM PST) (0400UTC) – 1. **1**	**2**	**3**	**4**	**5**	**RSGB** – 7-MHz Contest (phone) – 6-7. **FLORIDA STATE CONVENTION** – Miami, FL. Contact W4WYR – 6-7. **6**
NH-VT QSO PARTY – 7-9.** **ARRL HAMFEST** – Arlington Heights, IL. Contact WB9PWM – 7. **7**	**8**	**AMSAT** East Coast Net 3850 kHz 8PM EST (0100Z Wednesday Morning) **AMSAT** Mid-Continent Net 3850 kHz 8PM CST (0200Z Wednesday Morning) **AMSAT** West Coast Net 3850 kHz 7PM PST (0300Z Wednesday Morning) **9**	**W1AW QUALIFYING RUN** – 10. **10**	**11**	**12**	**OREGON QSO PARTY** – 13-15.** **YL-OM Contest** (phone) – 13-14.** **13**
14	**WEST COAST BULLETIN** – 9PM PDT (8PM PST) (0400UTC) 15. **15**	**AMSAT** East Coast Net 3850 kHz 8PM EST (0100Z Wednesday Morning) **AMSAT** Mid-Continent Net 3850 kHz 8PM CST (0200Z Wednesday Morning) **AMSAT** West Coast Net 3850 kHz 7PM PST (0300Z Wednesday Morning) **16**	**17**	**18**	**19**	**INTERNATIONAL DX CONTEST** – CW – 20-21. **20**
21	**22**	**AMSAT** East Coast Net 3850 kHz 8PM EST (0100Z Wednesday Morning) **AMSAT** Mid-Continent Net 3850 kHz 8PM CST (0200Z Wednesday Morning) **AMSAT** West Coast Net 3850 kHz 7PM PST (0300Z Wednesday Morning) **23**	**W1AW QUALIFYING RUN** – 24. **24**	**25**	**26**	**OHIO STATE CONVENTION** – Cincinnati, OH. Contact K8JE – 27-28. **27**
ARRL HAMFEST – Cuyahoga Falls ARC, Akron, OH. Contact Bill Sovinsky, 2305-24th St., Cuyahoga Falls, OH 44223 – 28. **ARRL HAMFEST** – Livonia ARC, Livonia, OH. Contact WA8GWL – 28. **ARRL HAMFEST** – Vienna Wireless Society, Vienna, VA. Contact K5RJ – 28. **28**						

**see coming events

FEBRUARY 13 AND 14; FEBRUARY 27, 28: YL-OM Contest. Phone starts Saturday, February 13, 1800 UTC, ends February 14, 1800 UTC. CW starts Saturday, February 27, 1800 UTC, ends February 28, 1800 UTC. All licensed men and women operators worldwide are invited to participate. OM's call "CQ YL" and YL's call "CQ OM." OPERATION: All bands may be used. No cross band operation. Nets/repeaters do not count. EXCHANGE: Station worked, QSO number, RS(T), ARRL section or country. Log entries must show time, band, date, and transmitter power. CW and Phone are separate contests. Submit separate logs for each contest. Awards: Cups/certificates. Send logs to: Sandra Heyn, WA6WZN, 962 Cheyenne Street, Costa Mesa, CA 92626 no later than April 5, 1982.

FEBRUARY 13 AND 14: Oregon QSO Party sponsored by Brand-X Amateur Radio Association from 1700Z Feb. 13 until 0800Z Feb. 14. From 1500Z Feb. 14 until 0000Z Feb. 15. Suggested frequencies: Phone — 1810, 3920, 7260, 14,300, 21,370, 28,600. CW — Send "CQ OR" — 60 kHz up from bottom of each band. Novice/Tech — 10 kHz up from bottom of each Novice/Tech band. Logs must be received by March 8, 1982. Mail to: Tim Burdick, WA7NVT, 138 N. 20th St., Philomath, OR 97370. SASE for results.

FEBRUARY 14. The Oregon Tualatin Valley Amateur Radio Club will operate a special event station, Sunday, February 14, commemorating the 123rd birthday of the State of Oregon, admitted to the Union on Valentine's Day in 1859. KA7CTP will operate from 9 AM to 6 PM PST on or near frequencies of 14.280, 21.360 and 28.510. An attractive certificate QSL will be awarded. Send 9 × 12 SASE or $1.00 to Marshall D. McKillip, 1175 NW 128th Street, Portland, Oregon 97229.

THE JUNIATA VALLEY AMATEUR RADIO CLUB'S 25th anniversary will be commemorated by a special event station operating at various times starting January 1. The club call is K3DNA, located in Lewistown, PA., Mifflin County. The station will operate on different bands, CW or phone. One contact with a club member entitles an operator to a club certificate.

FEBRUARY 5-14. The North Okanagan Radio Amateur Club along with the Vernon Winter Carnival Society are sponsoring a certificate to celebrate the 22nd annual Vernon Winter Carnival, Western Canada's largest. We will be operating daily from 2100-2400 Zulu and on Feb. 7 from 2000-0200 Zulu. Frequencies: 28.575, 21.375, 14.295 ± QRM. Send log information of 3 QSO's with Vernon area stations or one contact with club station (VE7NOR) to P.O. Box 1706, Vernon, BC, V1T 8C3.

More Details? CHECK—OFF Page 100

HAM MART

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ICOM IC3A/3AT 220-MHz handheld

ICOM announces a "second cousin" to the popular IC-2A series, the IC-3AT for the 220-MHz band. The IC-3AT is essentially identical in appearance, size, and features to the IC-2A series. All accessories, including battery packs, chargers, and microphone are completely compatible between the IC-2AT and IC-3A series, so accessories purchased for the IC-2A can also be used on the IC-3AT.

The IC-3AT also includes a sixteen-button Touchtone™ pad. It covers the entire 220-MHz band from 220-MHz to 224.99 MHz and is set up for both duplex and simplex operation. The power output is nominally 1.5 watts with the standard IC-BP3. The IC-3A system comes complete with IC-BP3 NiCd battery pack wall charger, belt clip, rubber duckie, and wrist strap. For more information, contact ICOM AMERICA, INC., 2112 116th Ave. N.E., Bellevue, Washington 98004.

Touchtone is a registered trademark of American Telephone and Telegraph Company.

TS-660 Quad Bander

Trio-Kenwood Communications announces a unique, new radio, the TS-660 "Quad Bander," an all-mode transceiver designed for operation on 6, 10, 12, and 15 meters. The TS-660 features built-in dual VFOs, a five-channel memory, memory scan, fm, SSB(USB), CW, an a-m operation, fluorescent digital frequency display, squelch, UP/DOWN pushbutton frequency control on the microphone, UP/DOWN pushbutton bandswitch, i-f shift, CW semi break-in with sidetone, S-meter, RIT control, and noise blanker.

The rf output power is 10 watts on SSB, CW, and fm, and 4 watts on a-m. The Quad Bander operates on 13.8 Vdc drawing 1 ampere in receive, 4 amperes in transmit. Additional information may be obtained by contacting Trio-Kenwood Communications, P.O. Box 7065, Compton, California 90224.

surge shunt

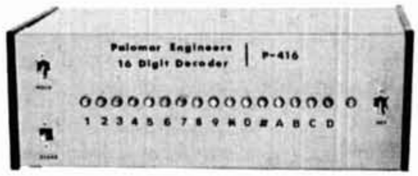
The R.L. Drake Company announces its new model 1549 Surge Shunt. The Surge Shunt protects solid-state communications equipment from damage by voltage transients entering the antenna system. These transients are usually caused by atmospheric static discharges or nearby lightning strikes.

The Surge Shunt can be used with receivers and with transceivers of up to 200 watts output. Convenient UHF-type coaxial connections are used, permitting use well into the UHF range. Price is $24.95.

For more information, contact R.L. Drake Company, 540 Richard Street, Miamisburg, Ohio 45342.

sixteen-digit DTMF decoder

New Model P-416 decoder converts incoming Touchtone™ signals to dc outputs on sixteen lines. Indica-



tor lamps show which digit is being received. Outputs will drive power relays directly and are either momentary or latched as selected by a panel switch.

Decoding is by digital logic with quartz crystal frequency control for long-term stability and operation over wide temperature variation. Operates on 115-volt ac power. Price is $385.

Write Palomar Engineers, 1520-G Industrial Ave., Escondido, California 92025 for further information.

12-ft antenna for 20/15/10

At last, a truly high performance short antenna designed specifically for hams with limited space. Where larger antennas require installation on costly towers or close to roof-top level, the lightweight and small DAP-31 makes this antenna ideally suited for mounting high and in the clear, using low-cost TV masts or tripods. Only the highest quality materials are used: 6061-T6 aluminum alloy for the element and capacitive-hat frames; fiberglass rod center insulator; aluminum or stainless steel hardware for all electrical connections; ultra-lowloss, air-wound, large-diameter inductor wound with 1/4-inch aluminum tubing.

The model DAP-31 sells for $165 with balun. For more information, contact Diversified Antenna Products, 2330 W. National Ave., Milwaukee, Wisconsin 53204.

memory keyer

Heathkit's SA-5010 uMatic Keyer uses a custom microprocessor to provide up to ten buffers for storing up to 240 characters of text or commands. These variable-length buffers eliminate wasted memory space by letting the user store text in several buffers and then string them together in any sequence. Command strings can also select the speed, weight, spacing, and auto-repeat count for each message so selected.

The SA-5010 employs a 20-position keypad for entries, and features easy-to-use integral capacitive touch paddles. A rear panel jack is provided for use of a mechanical paddle with the SA-5010, if so desired. A practice mode sends random code groups of random length and selectable types. The 100 different random sequences are repeatable, so the ham can check copy for accuracy. All 100 sequences are altered each time the keyer is turned on, said to give a total of 6,400 different practice sessions. Each sequence sends approximately 3,000 characters before repeating. The user can choose any speed between 1 and 99 words per minute, and any of 11 weight settings.

Text may be manually inserted into a buffer message being sent. A pause may be stored into text or command strings to cause the keyer to pause automatically for manual insertion of, for example, a station's RST report. The SA-5010 allows entry of text at whatever speed and weight are comfortable for the user. It can be sent at any other settings desired.

NEW products

CMOS memory with battery back-up retains the buffer contents, as well as the last-selected speed, spacing, weight, and repeat count whenever the keyer is turned off or unplugged. Built-in diagnostics check the microprocessor each time the keyer is turned on, and also test buffer memory whenever the uMatic Memory Keyer is reset. For the left-handed CW operator, a special two-key function reverses the paddles. And the keyer even remembers to turn itself off if the user forgets.

Mail order priced at $97.95, FOB Benton Harbor, Michigan, the SA-5010 features built-in side tone oscillator and speaker with variable pitch and volume controls. Phone jack and ear phone are included for private listening. A plastic case covers the die-cast zinc base, which is weighted to reduce movement during keying. The keyer requires the optional 120 Vac Heathkit PS-5012 power supply. Step-by-step instructions make this Heathkit project a two-evening kit. Write Heath Company, Dept. 350-115, Benton Harbor, Michigan 49022.

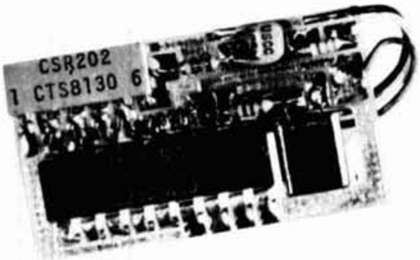
2-kW antenna tuner

Ten-Tec announces a new 2-kW antenna tuner/SWR bridge/power meter. The new tuner uses a reversible "L" configuration with a silver-plated roller inductor, high-voltage variable capacitor, and selectable fixed capacitors for greater versatility in impedance matching. The design automatically provides a low-*Q*, minimum-loss path when properly adjusted. Power ratings are 2 kW PEP, and 1 kW CW. Frequency range is 1.8 to 30 MHz. Model 229 matches conventional 50-ohm unbalanced outputs of transceivers or linear amplifiers to a variety of balanced or unbalanced load impedances. Antennas such as dipoles, inverted Vs, long random wires, Windoms, beams, rhombics, mobile whips, Zepps, Hertz, and similar types can be matched. A built-in balun converts one antenna to a balanced configuration if desired.

The built-in SWR bridge and dual-range power meter indicates SWR from 1:1 to 5:1 and power from 10 to 2000 watts. Model 229 net price is $249.00. For full information, write Ten-Tec, Highway 411 East, Seiverville, Tennessee 37862.

CTCSS encoder

Communications Specialists introduce the new SS-32M micro-miniature programmable CTCSS encoder for use in the Icom IC2AT handheld. The unit is based on the popular SS-32 encoder and is programmable



using jumpers. Measuring just 1.45 × 0.8 × 0.13 inches, the SS-32M may also be used in other applications where size is critical. Priced at $29.95.

A catalog is available on request. For more information, write Communications Specialists, Inc., 426 West Taft Avenue, Orange, California 92667.

Astatic's MK-I microphone

The Astatic MK-I Mobile King transistorized dynamic noise-cancelling microphone provides clear transmission for close talking. The MK-I has frequency response tailored for maximum voice clarity with 360-degree minimization of background noise.

NEW products

The handheld microphone is designed for CB, Amateur Radio, and SSB communications. A soft lip guard ensures proper close talk spacing.

The MK-I has a conveniently positioned press-to-talk lever and external slide controls for volume and tone. The controls permit easy accessibility for correct modulation level and tone quality. Power is supplied by a 9-volt

battery. The solid-state amplifier provides output impedance for compatibility with inputs of most transceivers. The six-wire coil cord allows virtually any type of transceiver hookup, including relay and electronic.

Maximum output level is −30 dB below 1 volt/microbar at 1 kHz into 1-megohm load. Output adjustment is 40 dB minimum range. Impedance: 5000 ohms maximum.

For more information contact The Astatic Corporation, Conneaut, Ohio 44030.

40-meter monobanders

KLM introduces two new monobanders for 40 meters. The two-element 7.2-2 and three-element 7.2-3 offer substantial gain and F/B coupled with modest size. Their design and performance characteristics are based on KLM's well known four-element "Big Sticker" but require considerably less airspace, and are usable with most standard ham rotators. Lossless linear-loading keeps element length to 46 feet and permits stacking within 4 feet of 20 meter beams.

The 7.2-2 sits on a 16-foot boom and weighs only 45 pounds. The 7.2-3 has a 32-foot boom, supported by stainless steel overhead guy cables and weighs 70 pounds. Both employ the same materials and construction as all of KLM's full size monobanders: 6063-T832 aluminum alloy elements and boom (3 inch O.D. × 0.065 inch wall), Lexan insulators, and all stainless steel hardware (except U-bolts). Each is supplied with KLM's 1:1 4-kW PEP ferrite balun.

For more information on these and other KLM antennas contact: KLM, P.O. Box 816, Morgan Hill, California 95037.

Hamtronics® repeater modules

Hamtronics, Inc., long noted for fine quality fm transmitters and receivers commonly used for building repeaters, has now completed its line of repeater modules by offering inexpensive COR and CWID modules.

The 3-inch-square COR module kit ($29.95) contains an electronic relay to actuate the transmitter when the receiver squelch opens. Adjustable tail and time-out timers are provided on the board as well as an audio mixer to combine the ID tone with the receiver audio for application to the transmitter. Another nice feature is a separate speaker driver amplifier stage which allows a local speaker to be operated completely independently from the repeater audio level setup control, without using hard-to-find L-pads.

The 3 × 7 inch CWID module kit ($59.95) contains a tone generator controlled by a 158-bit diode matrix. Adjustments on the board control the

NEW products

ID speed, tone, and time. A special output filter eliminates the key clicks found on many repeater ID'ers. Although many articles have appeared recently on PROM ID'ers, a survey by Hamtronics indicates that most still prefer a diode matrix because it can be easily reprogrammed at any time without further expense. The CWID module is constructed on a double-sided PC board with plated-through holes for easy assembly.

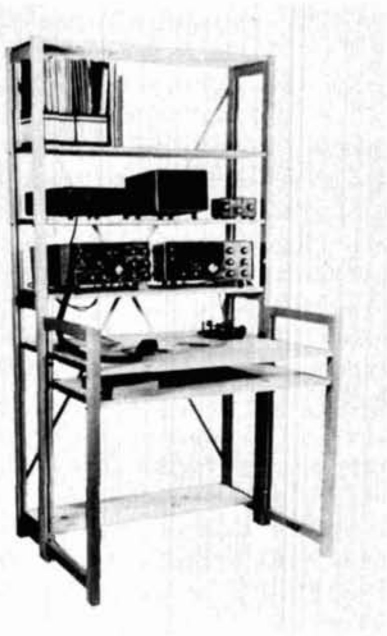
The CWID and COR modules are designed to be used with Hamtronics® transmitter and receiver modules as well as most other types.

For further information, contact Hamtronics, Inc., 65F Moul Road, Hilton, New York 14468, or phone 716-392-9430. A complete 1981 catalog, which includes all Hamtronics® fm and SSB equipment, is yours for the asking. (For overseas mailing, please send $2.00, or five IRCs.)

operation desk

Ricker Equipment, Inc., of Fort Wayne, Indiana, has introduced an operating desk for use with ham radio equipment and computers, called Shack Desk. It makes use of the space above desk-top level better than any other unit on the market. Shack Desk is made of wood — prefabricated uprights and lumber shelves. The shelves are edge-glued, designed to carry the weight of ham gear including linear amplifiers. The standard Shack Desk provides a desk top 30 inches deep by 36 inches wide with four shelves above the desk top level that are 12 inches deep and 36 inches wide. A second 18-inch-deep shelf 36 inches wide under the desk-top level provides handy storage for reference material. The unit is open-construction at the sides and back for maximum air circulation. The user may enclose the ends or back if desired, with lumber or paneling readily available.

For hams with plenty of space, there are larger Shack Desk units available. Shack Desk is sold unfinished, with the particular finish left up to the individual user. For further information, contact Ricker Equipment, Inc., P.O. Box 12304, Fort Wayne, Indiana 46863.



Volume 4: Disassembled Handbook for TRS-80

Volume 4: Disassembled Handbook for TRS-80 is the title of a newly released volume by Richcraft Engineering Ltd. that presents Morse code, Baudot, and ASCII programs for the Model I and Model III TRS-80 microcomputers.

Volumes 1 and 2 of the *Disassembled Handbook* series introduce the newcomer to assembly-language programming to the use of multifaceted CALLs in Level 2 ROM. As it says in the foreword to volume 4, these CALLs can "often save the student 1 to 2 years (of) study, and hours/pages of programming effort for most any program written in assembler." Volume 3 of the series is system oriented, and includes programs and hardware for implementing spooling, interrupts, interfacing to the outside world, A/D converters, D/A converters, and RS-232C adapter applications. Volume 4 presumes a working knowledge of Level 2 ROM.

Volume 4 of the series is aimed at radio communications. The material is presented in a format that will work with most any Model I TRS-80 with Level 2 BASIC and at least 16K of memory, *without* an expansion interface and RS-232C adapter. The programs will work with or without disk. Each chapter includes the modifications necessary to enable Model III users to make the minor changes necessary to ensure that all the programs in this volume will work as well on their machine as on the Model I.

There are ten chapters in Volume 4, dealing with the following subjects:

Chapter 1: Converting parallel ASCII from the keyboard to serial Morse output via the cassette output line. Code speed may be selected from 8-800 words/min.

Chapter 2: The dots, dashes, and spaces of Morse code are divided down into finite segments and the keyboard scanned during the element's output period.

Chapter 3: Converting serial Morse code input via the CASSIN line from the receiver's speaker to parallel ASCII presented to you on the video display.

Chapter 4: Combining the Morse transmit and receive programs into the single working entity. The CLEAR key serves as the operator's TRANSMIT/RECEIVE switch.

Chapter 5: Converting parallel ASCII input from the keyboard into serial Baudot radio teletype signals. Speeds of 60, 66, 75, and 100 words/min. may be selected.

Chapter 6: Converting serial Baudot input from the station's receiver to parallel ASCII that is presented to the operator on the video display.

Chapter 7: Combining the Baudot transmit and receive programs into a working entity. Twenty-two prepared messages may be selected and/or input a prepared message via keyboard.

Chapter 8: Converting parallel ASCII input from the keyboard into serial 110-Baud ASCII radio teletype signals.

Chapter 9: Converting 110-Baud serial ASCII radio teletype from the station receiver into parallel ASCII that is presented to the operator on the video display.

Chapter 10: Combining the ASCII transmit and receive programs into a working entity. Twenty-two prepared messages may be selected and/or input a prepared message via keyboard.

Volume 4: Disassembled Handbook for TRS-80 is written in a clear, question-and-answer format. It is available from Richcraft Engineering Ltd., #1 Wahmeda Industrial Park, Chautauqua, New York 14722.

tower trailer

For those special situations that require mobility, Aluma Tower Company introduces an all-steel trailer for transporting and erecting any Aluma aluminum or steel tower. Ideal for mobile demonstration work, civil defense, remote signal testing, and many other situations, the tower trailer combination is easily towed. Once in place, the tower is tilted up and cranked into position. The trailer acts as a secure base.

For more information, contact Aluma Tower Company, 1639 Old Dixie Highway, Box 2806, Vero Beach, Florida 32960.

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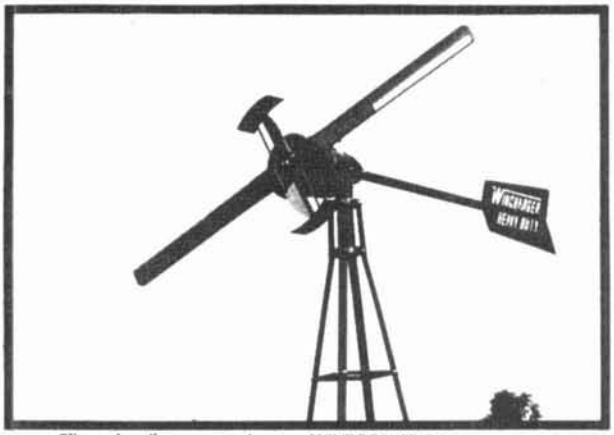

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